A STUDY OF DELAY LINES AND THE DEVELOPMENT OF DESIGN CRITERIA FOR HF COMMUNICATIONS RECEIVER NOISE BLANKERS

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THESIS

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ABSTRACT

Atmospheric and man-made disturbances which result in noise pulses of random amplitude and randomly spaced in time have long been experienced in the communications field. Design features of an RF blanker for this type of noise are investigated. The specifications of functional blocks are established.

One of the functional blocks of an RF blanker is a linear delay line. The main features and design methods for passive delay lines are investigated. An active RC circuit to realize the delay function is proposed. This network and another are constructed and their characteristics obtained.

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I. INTRODUCTION

A. NEED FOR STUDY

The problem of impulse noise has long been experienced by people in the radio communications field, especially with mobile communications equipment [1]. The type of noise referred to here consists of impulses randomly spaced in time and generally of random amplitude. Atmospheric and man-made disturbances are the primary sources of this type noise. Distinct among these is the impulsive noise produced by the automobile ignition system. This type of noise is characteristically composed of pulses of very short duration and very fast rise time, occuring at a nearly periodic rate which is dependent upon the engine speed.

In general the impulse noise frequently has an amplitude in the millivolts range, sometimes as high as several volts. The duration of the pulses, however, seldom exceeds 25 to 50 microseconds [2].

When impulse noise is induced in the antenna of a receiver, it produces two effects [3]. One, it excites the front end of the receiver to produce a burst of RF energy at the resonant frequency of the tuned circuits, which then proceeds through the succeeding stages of RF and IF amplifiers as if it were a signal. Second, the tuned circuits at the front end will produce damped oscillations after the pulse has ended. The ringing duration is inversely proportional to the bandwidth of the circuit. Thus, in a modern communications

receiver with high selectivity, an impulse that may last only a fraction of a microsecond at the antenna produces a ringing signal that may be as long as 300 microseconds in the latter part of the receiver.

Many systems have been developed to reduce or eliminate this noise. However, they are either ineffective and require constant operator adjustments, or they are large, cumbersome and expensive. A need exists for a simple, low cost system of wide dynamic range that will remove this noise without degrading the desired signal.

1. Description of RF/RF noise blanker

One method of reducing the impulsive noise interference is to allow the noise to progress through the receiver and then, at a later stage, to limit the amplitude of the noise to that of the desired signal. This method is referred to as noise limiting or clipping.

Another method utilizes an electronic switch in series with the signal path, which renders the receiver inoperative during the duration of the noise pulse. This method is called silencing or blanking.

It is desirable to remove the noise at as early a stage as possible. Theoretically, one would like to have it removed right at the antenna input point, which suggests the use of an RF/RF noise blanker. The term RF/RF comes from the noise being detected at the RF frequencies and being removed at the RF frequencies, before mixing. A general scheme of an RF/RF blanker is shown in Figure 1.

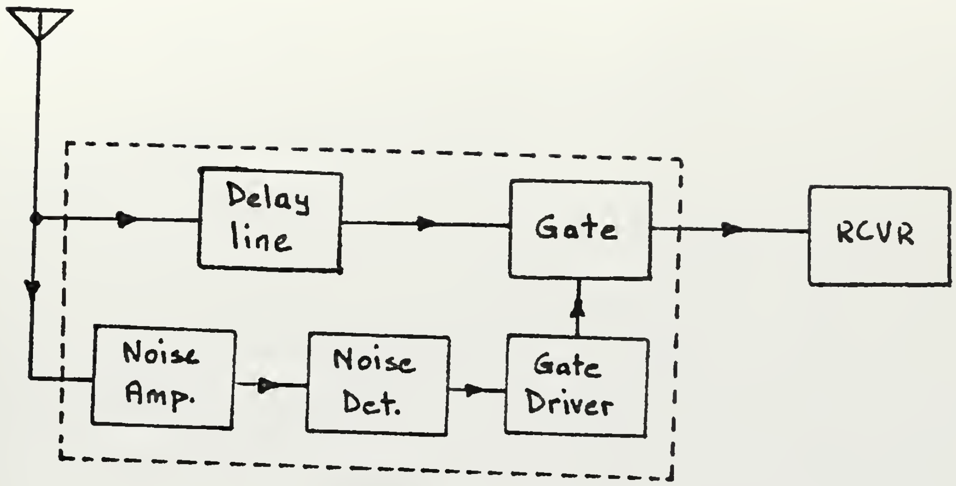


Figure 1. RF/RF noise blanker

In order that noise pulses be eliminated completely, it is necessary that the controlled stage, or gate, be turned off prior to the arrival of the leading edge of the noise pulse. One hundred to three hundred nanoseconds are necessary to recognize the noise pulse and produce the gate turn-off pulse, which requires that the signal be delayed by an amount of time greater than these combined times [4]. One channel in Figure 1 is the RF signal path which provides necessary delay to the noise pulse so that gate pulse and noise pulse coincide. The other channel is the noise channel, where the noise pulse is amplified and detected for triggering the gate driver circuits. The gate circuit opens the signal path for a predetermined fixed interval after the gate driver delivers the blanking pulse. Therefore, holes are "punched" in the signal and information is lost for that interval. The duration of the blanking interval is much shorter than the

highest information frequency period, therefore, unless the repetition rate of the noise pulses is high, compared to the highest information frequency, blanking is unnoticable.

B. STATEMENT OF THE PROBLEM

The purpose of this study is to establish the feasibility of components such as the noise amplifier, detector and delay line for a low-cost RF/RF noise blanker for use with HF communications receivers. Two types are considered:

1) A simple, inexpensive, add-on device that will require no modifications to available receivers.
2) A built-in blanker which will be an integral part of a receiver.

Noise amplifier and detector circuits for both types of blankers are investigated and a tuned amplifier for noise detection is constructed and tested.

Passive and active linear delay lines are investigated; for active delay lines, simple feasibility models are constructed and tested.

C. REVIEW OF RELATED WORK

The military does not routinely use noise blankers on HF communications receivers, largely due to the complexity and the high cost of equipment to satisfy military requirements. Commercial use of blankers on the market is also very limited. Manufacturers offer blankers as optional accessories and these blankers are only available with the highest priced equipment [5]. In recent years a variety of noise blankers have

been available for amateur radio receivers, although almost all of them are IF/IF or RF/IF types. These blankers, however, are offered as accessories at high cost.

Two feasibility studies have been performed under military contract for RF/RF blanker.

1) The Lightning and Transients Research Institute developed, under ONR sponsorship, a blanker optimized to eliminate precipitation static in aircraft ADF receivers [6]. Development was completed in 1956 and the unit does not use present day components and/or techniques.

2) Southwest Research Institute manufactured for the Department of the Air Force a blanker, called by them a "VHF Time Domain Filter," which was intended for use in the 30-300 MHz frequency range [7]. Also, recently the same institute manufactured an HF version of the above blanker. Solid state devices were extensively used in both blankers. However, equipment was large, complex, needed many operator adjustments and costs $40,000.

The only other reference to a simple blanker of RF/RF type in the open literature seems to be "Elimination of impulsive noise interference from communications receivers by the RF blanker method," by W.A. Norman, an M.S. in Electrical Engineering thesis at the Naval Postgraduate School [4]. In this study, recommendations were made for the performance characteristics of functional blocks and an experimental

feasibility model was constructed. This experimental model achieved 60 db improvement in the noise elimination and proved the superiority of the RF/RF blanker to the others. This blanker used a coaxial cable as a delay line, was relatively narrow band (up to 14 MHz) and assumed no strong CW signals in the HF spectrum.

II. PERFORMANCE REQUIREMENTS OF NOISE BLANKER CIRCUITS

A. NOISE AMPLIFIER AND DETECTOR

Although noise pulses at the antenna may be quite large compared to the desired signal, they are much too small to trigger a noise detector, which is preset to a minimum input signal of approximately 100 millivolts. For the blanker to operate on noise pulses of a few microvolts, the noise amplifier must have a gain of about 100 db. To prevent overloading on large noise pulses, the amplifier must either be allowed to saturate or a detector-derived fast AGC be included to obtain a dynamic range over 80 db. As the antenna sees two channels in parallel, the front end of the noise amplifier must have an input impedance over the receiving range to the receiver, typically from two to thirty MHz, in order not to reduce significantly the sensitivity of the signal channel.

If there were no strong undesired signals in the HF spectrum, a wideband amplifier of 20 MHz bandwidth could be used to amplify the noise pulses prior to triggering the gate driver circuits. Normally very strong signals will appear in the HF spectrum and could be detected as if they were

noise pulses, thus degrading the usefulness of the blanker. To prevent this, two different solutions are possible.

1) An RF pre-amplifier, tuned to the desired frequency replaces the video amplifier. The output of this amplifier is fed to the two channels, the gate driver and the delay line. A block diagram is shown in Figure 2.a. In this manner the undesired strong signals can be eliminated. Also an AGC voltage can be derived from the desired signal and applied. This provides a means of comparison of signal and the noise pulse. There are two drawbacks to this system.

    a) If the total Q of the tuned amplifier is greater than 10-20, the noise pulses will cause ringing in the tuned circuits and will be stretched. This will require long blanking times. On the other hand, if the Q is lower than 10-20, strong undesired signals with frequencies close to the desired frequency will be able to trigger the gate driver, hence, degrade blanker operation.

    b) This system will require operator adjustments, unless it is an integral part of a receiver.

2) Looking at the spectrum of a typical noise pulse, one sees that it is "white" up to 300 MHz. This characteristic can be used to advantage to detect the noise pulse. Three or more band-pass amplifiers with band widths of 500 KHz or so, centered on frequencies evenly spaced in the HF spectrum will provide inputs

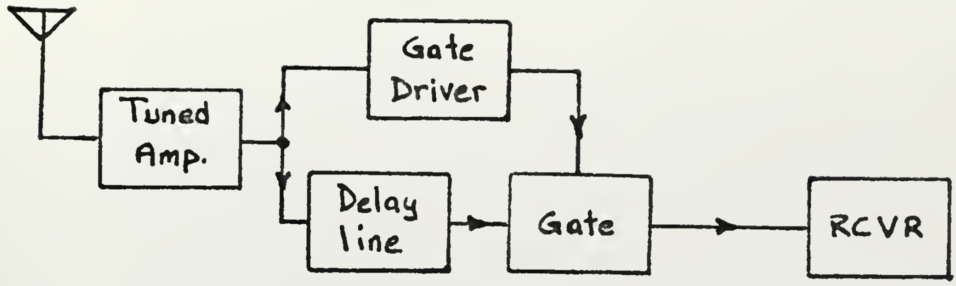


Figure 2.a. TRF blanker

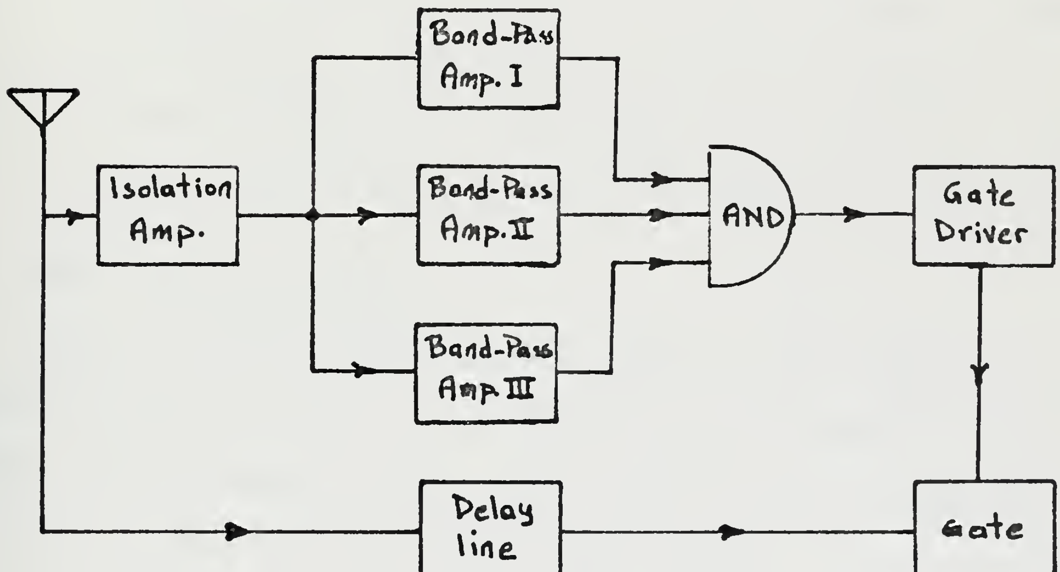


Figure 2.b. "And-gate" blanker

to an AND gate, which in turn detects time coincident pulses. Having three strong signals on the selected center frequencies of the above band-pass amplifiers is very unlikely. Therefore, fast rise time pulses will be detected while strong CW-RF signals will be eliminated. In Figure 2.b, a block diagram for this circuit is shown. The output of the AND gate then is fed to the gate driver. This system also suffers from narrow-band pulse stretching, but to a lesser extent than the first system described. An advantage is that it does not require operator adjustments.

An amplifier with center frequency of 13.85 MHz and $Q \cong 40$ (Figure 3), providing 17 db gain was constructed and its response to narrow pulses is shown in Figures 4.a and 4.b; the frequency response is shown in Figure 5. This amplifier can be used as a noise amplifier to provide an input to the AND gate.

B. GATE DRIVER AND GATE

The gate driver must operate whenever the output of the noise detector exceeds a threshold of approximately 100 millivolts. Once triggered, the output should be independent of the width and amplitude of the input pulse. A fast recovery, one-shot multivibrator may be used to meet these requirements.

The gate should have, ideally, infinite attenuation when it is closed and negligible insertion loss while open. Also, the gate must be capable of being rapidly turned on and off and must not generate any noise of its own. The RF signal

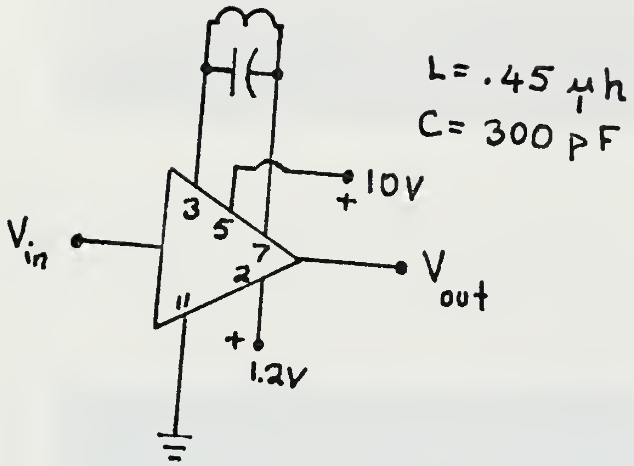


Figure 3. Tuned amplifier for noise channel using 1C, RCA CA3023

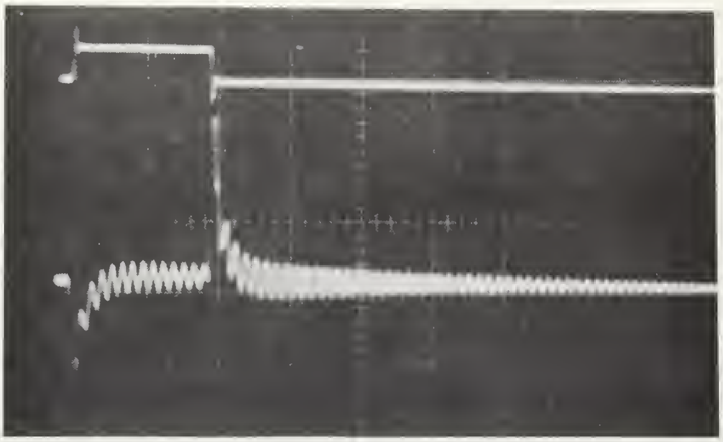


Figure 4.a. 1   sec pulse applied to the tuned amplifier. (Horizontal scale .5   sec/cm)

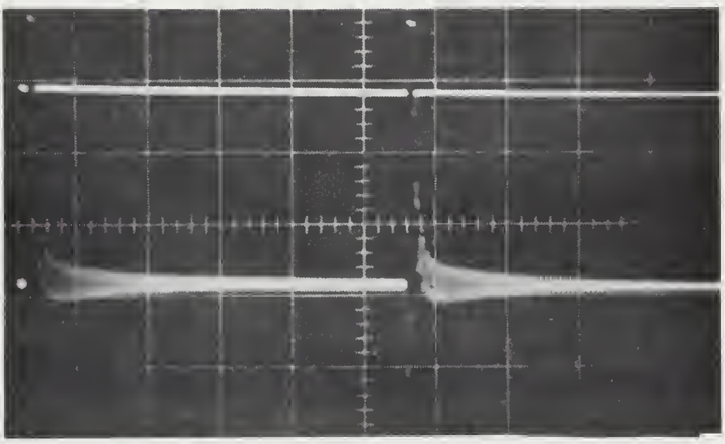


Figure 4.b. 1   sec pulse applied to the tuned amplifier. (Horizontal scale 1   sec/cm)

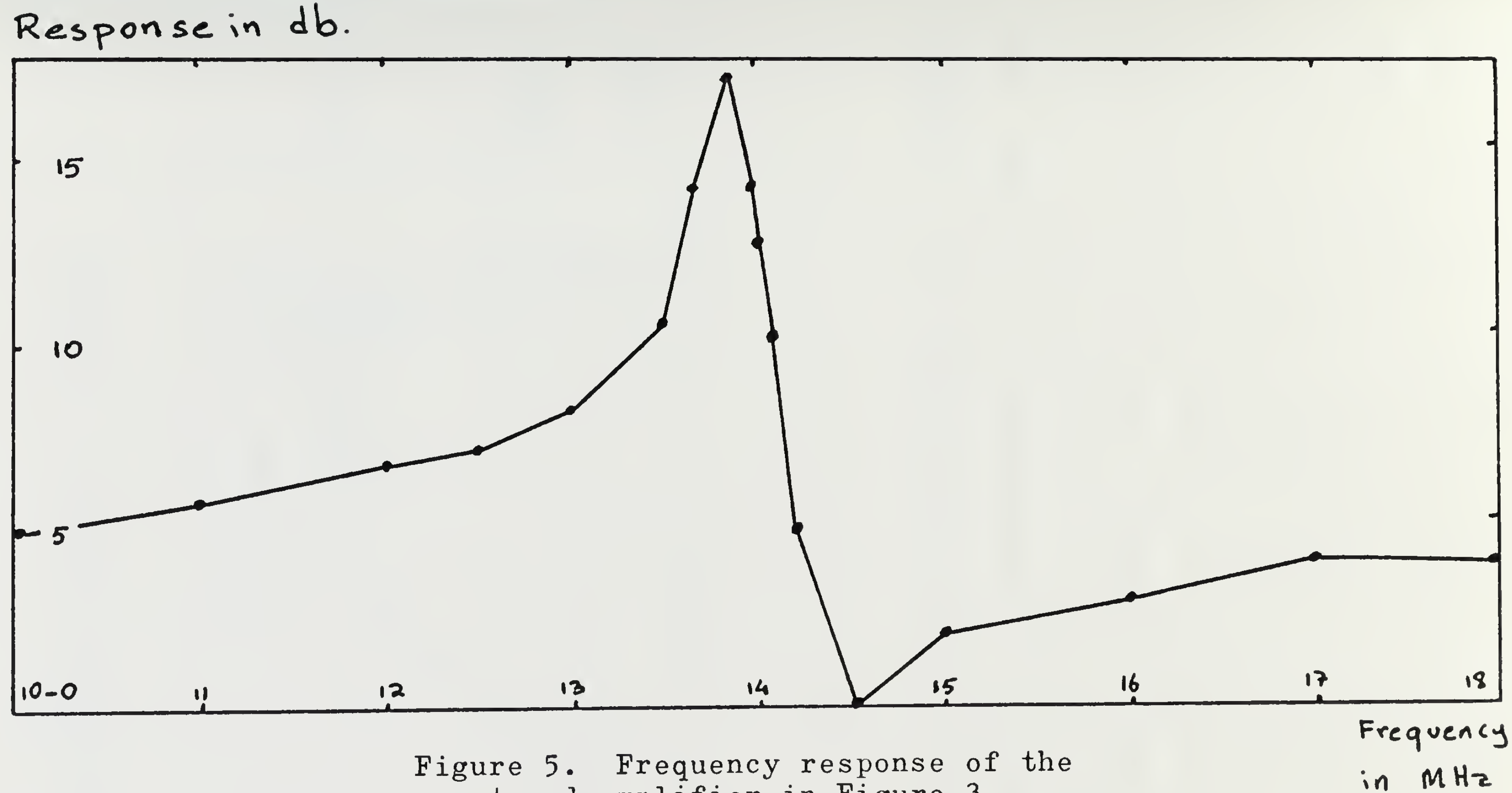


Figure 5. Frequency response of the tuned amplifier in Figure 3

level at the antenna may be as low as a fraction of a microvolt. Any disturbances produced by the gate must be well below that level and this requirement establishes turn-on and turn-off speeds for the gate. A smooth, rounded pulse must be produced as the gating pulse. It should have a rise time of 100 to 200 nanoseconds. This rise time is the majority of the delay time required of the delay line. The gate driver provides this pulse. Ringing and ripple on that pulse can be tolerated, since the gate will be off during that time, but ringing at the trailing end of the pulse is not acceptable. A sample of this pulse shape is given in Figure 6.

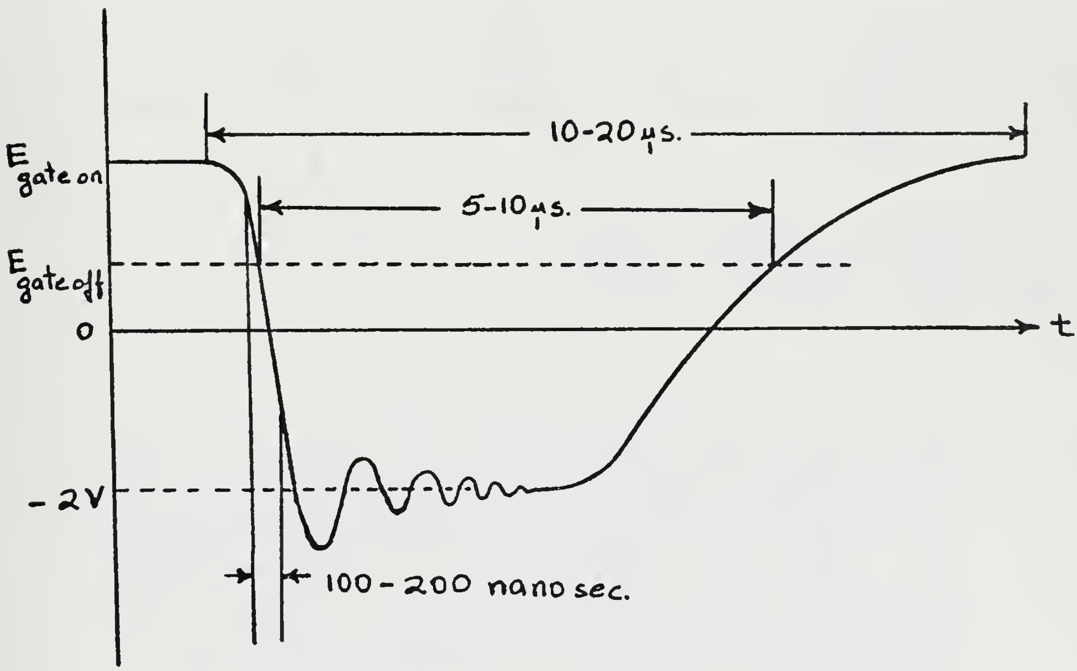


Figure 6. A typical gating pulse

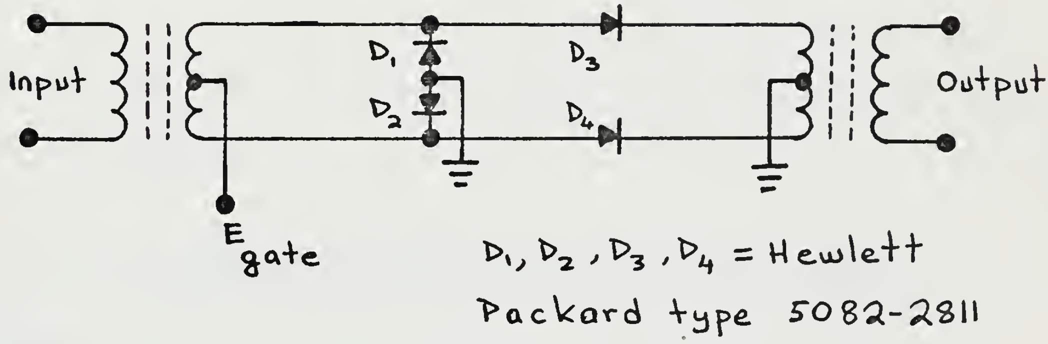
Although there are several gate arrangements possible for this application, because of the reasons below, a four diode gate using hot-carrier devices is recommended. The reasons are:

1) Simplicity and reliability.

2) Reduction in isolation at higher frequencies due to capacitance is minimized.

3) Low insertion loss and high isolation.

A schematic diagram and performance characteristics for this gate are given in Figure 7. [4].



Frequency	14 MHz.	21 MHz.	28 MHz.
Insertion loss	2db	2db	3db
Isolation	66db	63db	61db

Figure 7. 4-hot carrier diode gate and its performance characteristics

C. DELAY LINE

Since it takes a finite amount of time - approximately 100-200 nanoseconds - to develop a properly shaped gate pulse in the noise channel, it is necessary that the signal path be delayed at least this same amount of time, because the gate must be closed before the arrival of the noise pulse. This is accomplished by a delay line that has a cut-off frequency well beyond the highest frequency of interest. For an HF noise blanker this frequency will be 30 MHz. Therefore, the delay line must have negligible attenuation up to 30 MHz, low noise figure (less than 5 db) and low dispertion. A figure of merit for a delay line is the ratio of the delay time over rise time which can be related to the band-pass (width) of the line. For this application, this ratio must be greater than 20.

III. DESIGN AND MEASUREMENTS OF DELAY LINES

A. DESCRIPTION OF DELAY FUNCTION

The ideal characteristics required for a constant time-delay function (Maximally-flat delay form of response) is shown in Figure 8.a. To accomplish this,

$$G_{12}(s) = e^{-s} = \frac{1}{\cosh s + \sinh s} \quad \text{for T (delay time)}$$

normalized to $T = 1$ seconds. Then the approximation to this function due to L. Starch [8] is:

$$G_{12}(s) = \frac{\frac{1}{\sinh s}}{1 + \frac{\cosh s}{\sinh s}}$$

The series expantion of sinh s and cosh s is given by:

$$\sinh s = s + \frac{s^3}{3!} + \frac{s^5}{5!} + \frac{s^7}{7!} + \cdots$$

$$\cosh s = 1 + \frac{s^2}{2!} + \frac{s^4}{4!} + \frac{s^6}{6!} + \cdots$$

$$\frac{\cosh s}{\sinh s} = \coth s = \frac{1}{s} + \cfrac{1}{\frac{3}{s} + \cfrac{1}{\frac{5}{s} + \cfrac{1}{\frac{7}{s} + \cfrac{1}{\cdots}}}}$$

Let, $M = \cosh s$, $N = \sinh s$ and M_n and N_n be the truncated series expantion of cosh s and sinh s respectively. Then

$$G_{12}(s) = \frac{b_0}{M+N} \cong \frac{b_0}{M_n + N_n} \quad \text{where}$$

b_0 is a constant and $M_n + N_n = B_n(s)$ is a Hurwitz polynomial. The coefficients of $B_n(s)$ are given in Table 1 and the effect of truncation, in Figure 8.b.

The recursive formula to obtain the coefficients of $B_n(s)$ is $B_n = (2n-1)B_{n-1} + s^2 B_{n-2}$.

Step responses of truncated delay functions up to the fifth order are shown in Figures 9.a through 9.e, and magnitude and phase versus frequency responses for fifth order approximation are given in Figures 10.a and 10.b.

B. PASSIVE DELAY LINES

The delay function described above can be realised in many different ways, one of which is by passive networks. Those networks can be either distributed or lumped-constant element type.

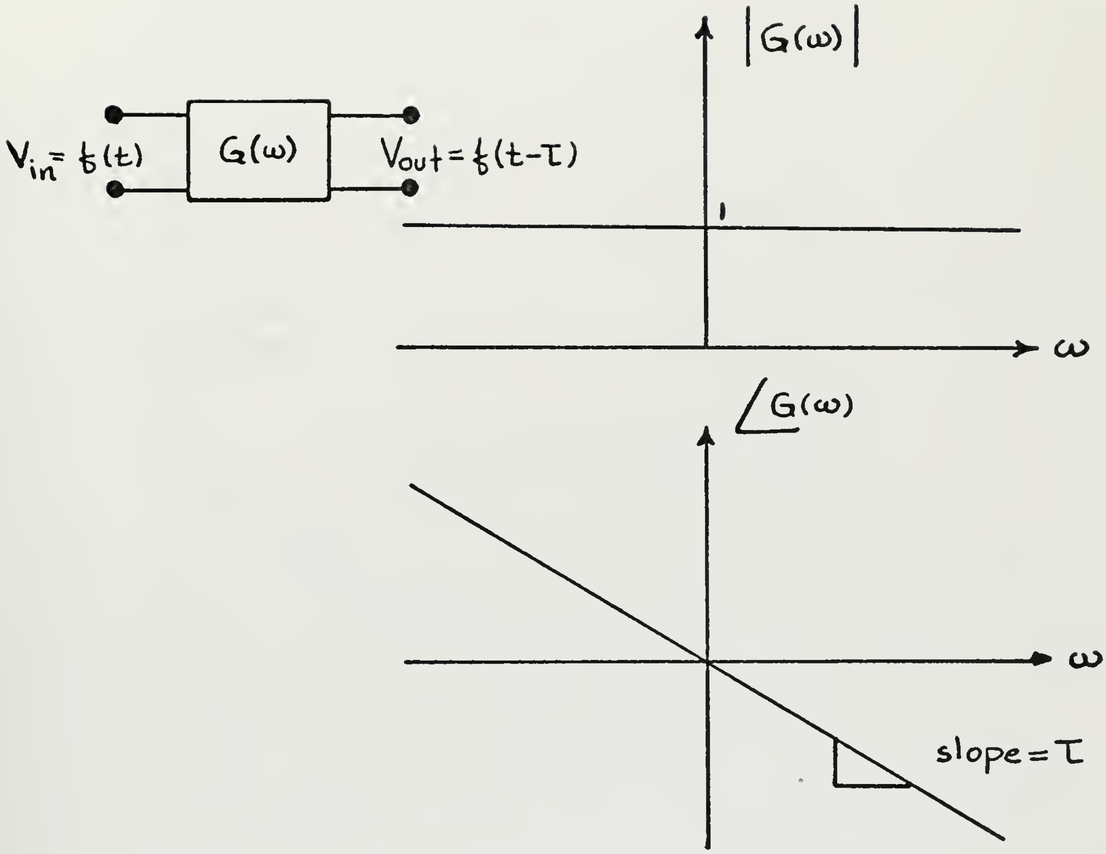


Figure 8.a. Maximally flat delay form of response function characteristics.

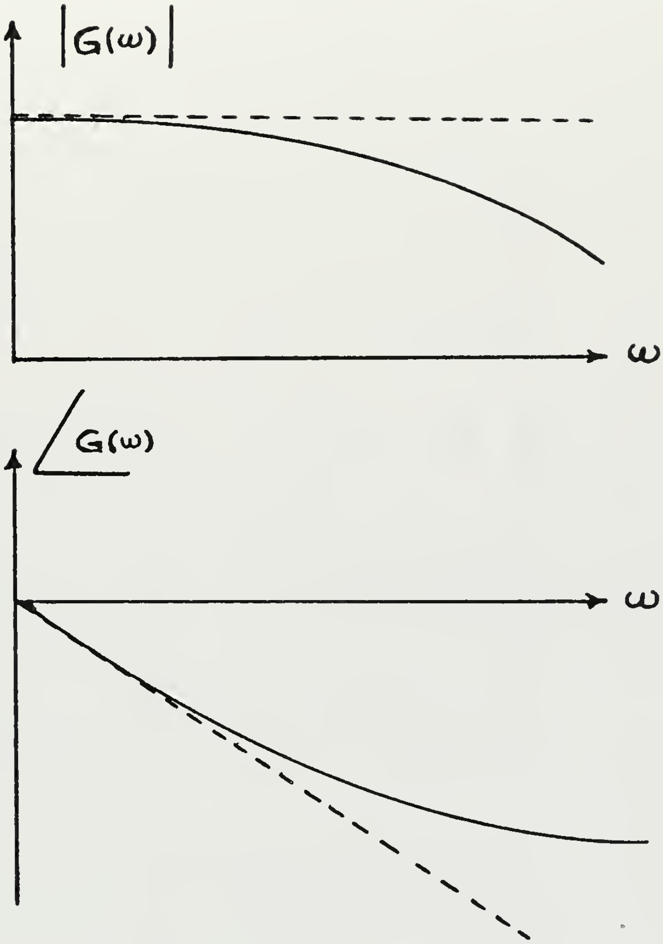
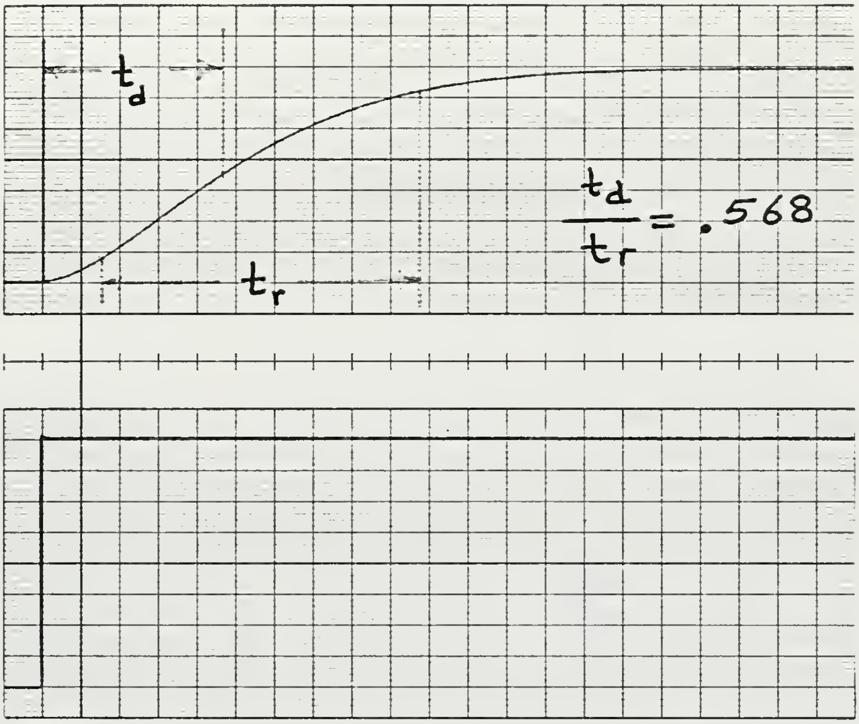


Figure 8.b. The effect of truncation on delay function.

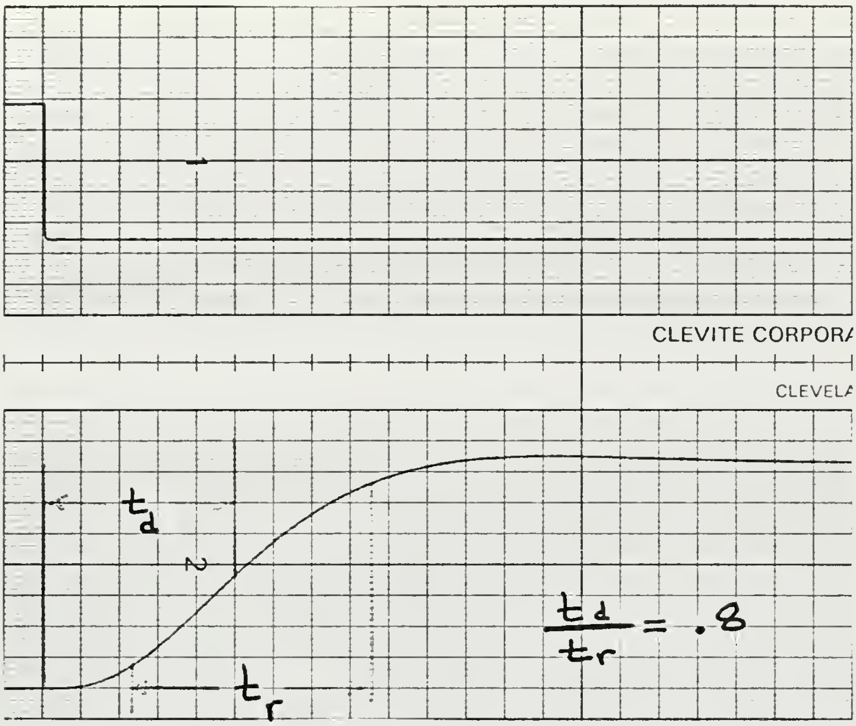
Table 1

Order	b_0	b_1	b_2	b_3	b_4	b_5	b_6	b_7
0	1							
1	1	1						
2	3	3	1					
3	15	15	6	1				
4	105	105	45	10	1			
5	945	945	420	105	15	1		
6	10,395	10,395	4,725	1,260	210	21	1	
7	135,135	135,135	62,135	17,325	3,150	378	28	1



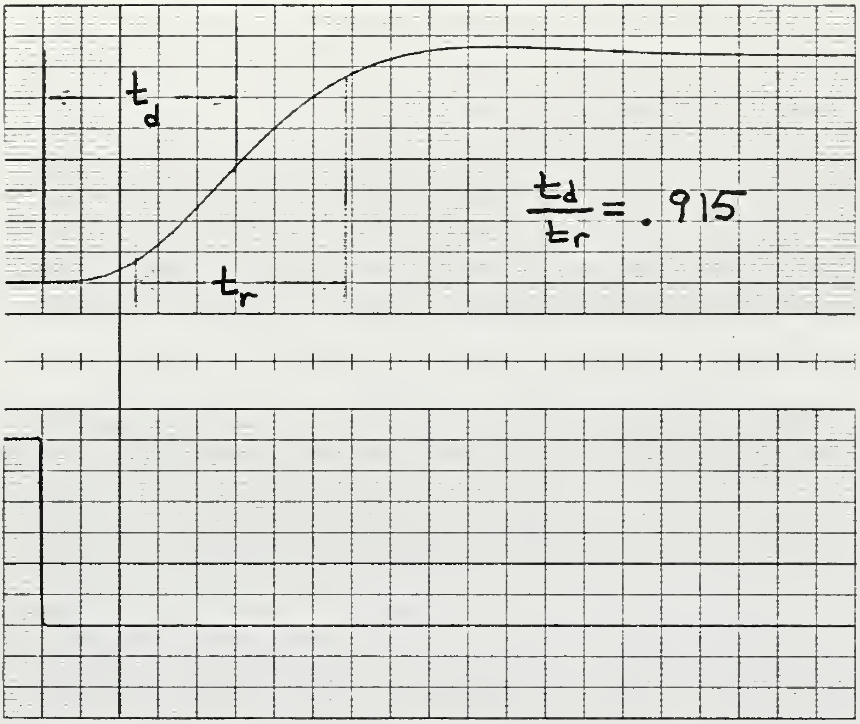
$$G(s) = \frac{H}{s^2 + 3s + 3}$$

Figure 9.a. Maximally flat delay function, truncated at second order. Response to a step input.



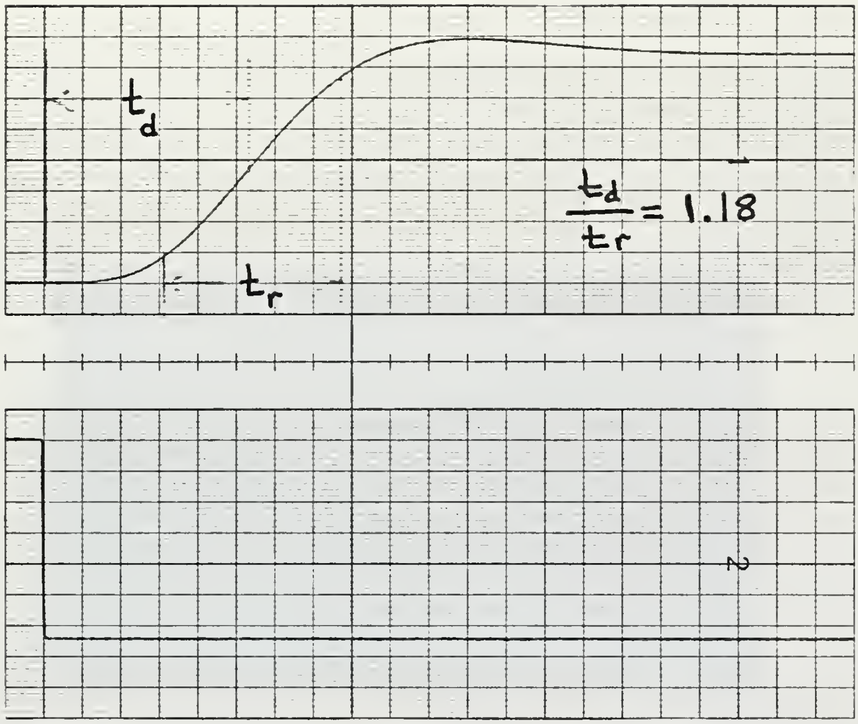
$$G(s) = \frac{H}{s^3 + 6s^2 + 15s + 15}$$

Figure 9.b. Maximally flat delay function, truncated at third order. Response to a step input.



$$G(s) = \frac{H}{s^4 + 10s^3 + 45s^2 + 105s + 105}$$

Figure 9.c. Maximally flat delay function, truncated at fourth order. Response to a step input.



$$G(s) = \frac{H}{s^5 + 15s^4 + 105s^3 + 420s^2 + 945s + 945}$$

Figure 9.d. Maximally flat delay function, truncated at fifth order. Response to a step input.

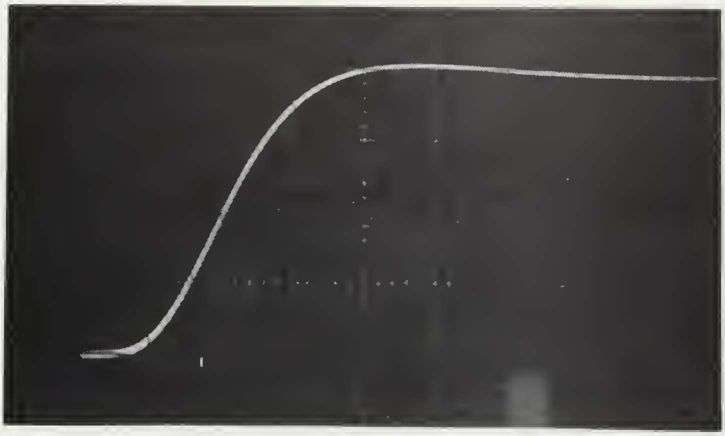


Figure 9.e. Maximally flat delay form of response, truncated at fifth order, response to a step imput. [Computer simulation of the transfer function time scaled T = 1 milliseconds] (Horizontal scale .5 milliseconds/cm)

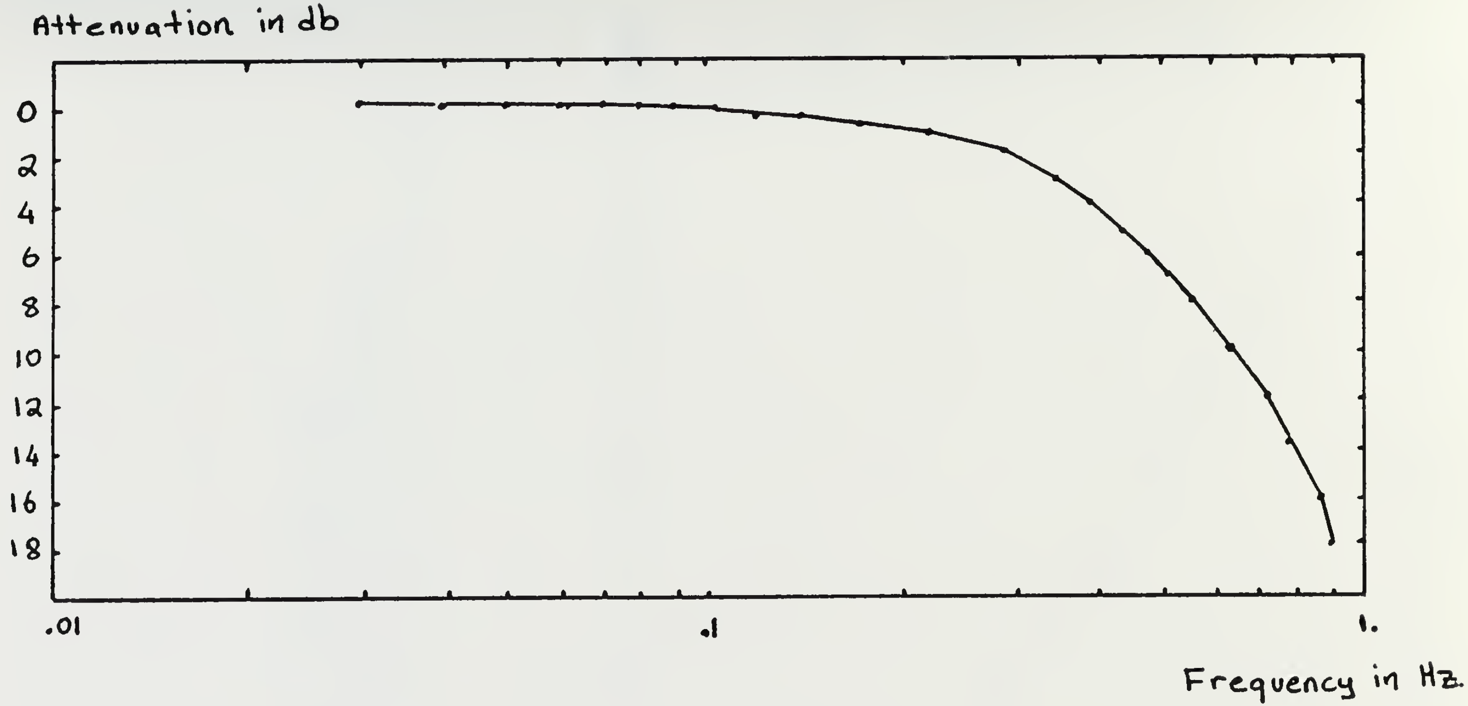


Figure 10.a. Maximally flat delay function truncated at fifth order, measured magnitude characteristics.

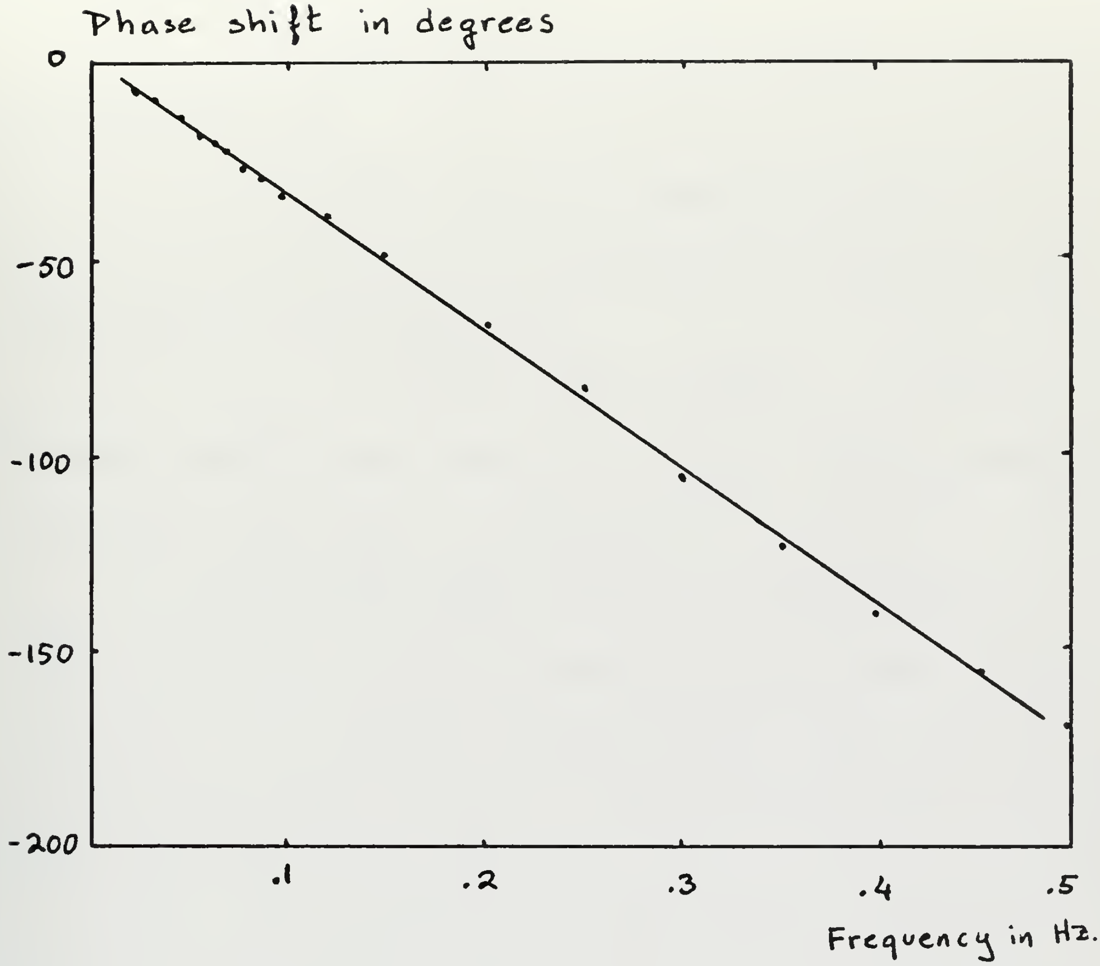


Figure 10.b. Maximally flat delay form of response, truncated at fifth order, measured phase characteristics. (T scaled for 1 sec.)

The main design parameters of an electromagnetic delay line are its impedance,

$$Z = \sqrt{\frac{L}{C}} \quad \text{(ohms, henries, farads)}$$

and its delay time,

$$T = \sqrt{LC} \quad \text{(seconds, henries, farads).}$$

For a given impedance and time delay, the inductance and capacitance are determined by,

$$L = TZ \quad \text{and} \quad C = T/Z.$$

The most general delay line is a coaxial cable of length "L." Its delay, regardless of its impedance, is [9]

$$T = \frac{1}{3}\, 10^{-10} L \sqrt{k} \quad \text{(seconds, centimeters)}$$

where "k" is the dielectric constant of the space between the conductors. For example, a polyethylene cable with k equal to 2.25 will yield a delay of 1 microseconds per 200 meters. This delay time is too small for practical delay-time design. To increase the delay time per unit length, one of the conductors is changed into a long, thin coil. This increases the inductance, hence the delay time per unit length. The characteristic impedance also increases, and a distributed delay line is obtained. These types of delay line have impedances from 200 to 3000 ohms.

For very low impedances or for obtaining longer delays in smaller physically sized delay lines, it is more convenient to lump the shunt capacitances, than to distribute them along the winding. Also, attenuation largely due to dielectric loses can be minimized by suitably chosen lumped capacitors. This leads to a lumped element type delay line.

1. Constant-k sections design

The delay in a simple low-pass filter network comprising series inductances and shunt capacitances rises appreciably with frequency. A lumped line is made up of a series of cascaded symmetrical networks such as shown in Figure 11.a. The "image or characteristic impedance Z_0" of this network is defined as, "if the network is terminated in Z_0, the impedance seen looking into the input terminals is also Z_0." Then,

$$Z_0 = \sqrt{\frac{L}{C}\left(1 - w^2 \frac{LC}{4}\right)}$$

and it can be shown that this equation reduces to $Z_0 = \sqrt{\frac{L}{C}}$ as previously mentioned, provided that,

$$f \ll \frac{1}{\pi\sqrt{LC}} \equiv fc$$

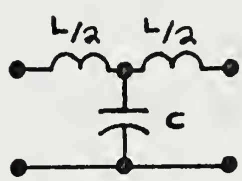


Figure 11.a A section of constant-k lumped element delay line.

This network has a third order all-pole transfer function and is called a "constant-k" delay line. If the Fourier spectrum of the input signal consists of frequencies much less than fc, there will be no distortion and the incoming signal will simply be delayed by a time

$$t_s \cong \sqrt{LC} \quad \text{(seconds, henries, farads)}$$

It has been found experimentally that the delay of a section terminated at both ends by Z_0 is

$$t_s = 1.07\sqrt{LC}$$

and the rise time (time between 10% and 90% points) is

$$t_{r_1} = 1.13\sqrt{LC}.$$

Also, "n" such sections cascaded will give

$$t_d = n t_s$$

$$t_{rn} = \sqrt[3]{n}\, t_{r1}$$

For a given t_d and t_r the number of sections required is given by

$$n = 1.1(td/tr)^{1.5}.$$

Then L and C is determined by

$$L = \frac{t_d \cdot Z_0}{1.07 . n} \qquad C = \frac{t_d}{1.07 . n . Z_0}.$$

In the noise blanker problem the required minimum delay time over rise time is 20. For 250 nanoseconds delay time and above ratio of 20, 95 sections are required with

$L = .492$ microhenries and

$C = 12.3$ picofarads.

This delay line can easily be built by using toraidal cores but will have unacceptable loss and high cost.

2. m-derived sections design

In the constant-k delay line, time delay is not constant over the pass band but, rather increases with increasing frequency. The dependancy of delay time on frequency can be lessened considerably by permitting some coupling between the two series inductors of the constant-k section;

therefore, m-derived sections are obtained. The optimized coefficient of coupling which will give the flattest time delay versus frequency plot is $k = .237$ [10]. The equations for n, t_s, L and C for given t_d, Z_0, t_d/t_r are,

$$t_s = 1.27\sqrt{LC}, \qquad n = .94\left(\frac{t_d}{t_r}\right)^{1.5},$$

$$L = .515\,\frac{t_d Z_0}{1.20n}, \qquad C = 1.27\,\frac{t_d}{1.20nZ_0}.$$

In Figure 11.b a section of this line is shown, indicating the sign of coupling.

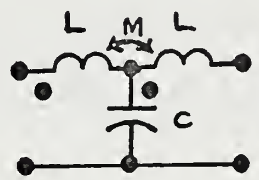


Figure 11.b. A section of m-derived, lumped element delay line.

The delay line with chatacteristics

$$t_d/t_r = 20, \quad t_d = 250 \text{ nanosecond}, \quad Z_0 = 200 \text{ ohms}$$

will require $n = 85$, $L = .252$ microhenries and $C = 15.6$ picofarads. Note that the number of sections has decreased from 95 for constant-k sections to 85 m-derived sections, therefore, a "less lossy line" is obtained. M-derived sections have a magnitude undershoot of 12 percent. Step responses of an m-derived section and a fifth order approximation to maximally flat delay function are given in Figures 12.a and 12.b, clearly indicating the undershoot of m-derived section.

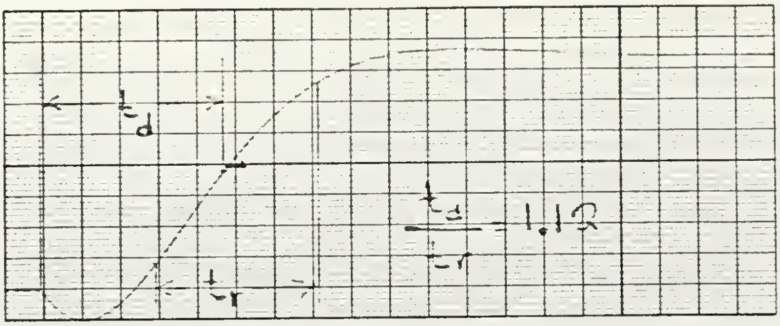


Figure 12.a. m-derived section

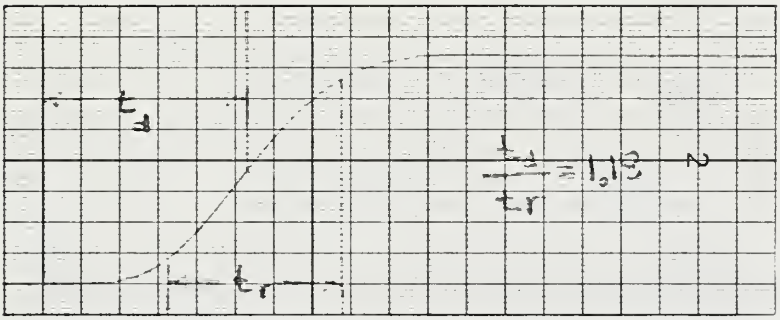


Figure 12.b. Fifth order approximation of maximally flat delay function.

Figures 12a,b. Comparison of responses of, M-derived section and fifth order approximation of maximally flat delay function, to step input.

Construction of an m-derived delay line by using toroidal coils needs special attention. Two coils will necessarily by wound on a single toroidal core to obtain coupling, but the coefficient of coupling can not be controlled and is very close to unity. To obtain similar characteristics to that of an m-derived section where the coefficient of coupling is optimized to $k = .237$, the section is made unsymmetrical by making the inductances in series, unequal. If the first inductance is L_1 and the second L_2, the equations become,

$$L_1 + L_2 = 1.3 \frac{t_d \ Z_0}{1.20 \ n} \quad \text{and}$$

$$(L_1 . L_2)^{1/2} = .122 \frac{t_d Z_0}{1.20 \ n}.$$

The delay line for the noise blanker previously calculated for a symmetrical m-derived delay line will require $L_1 = .496$ microhenries, $L_2 = .007$ microhenries, when constructed with toroidal cores.

C. ACTIVE LINEAR DELAY LINES

To overcome the losses in passive delay lines, active networks can be used to realize transfer functions with linear phase versus frequency characteristics. The simplest of those transfer functions is the second order approximation to the maximally-flat delay function.

$$G_{12}(s) = \frac{H}{s^2 + 3s + 3}$$

The network shown in Figure 13 has a transfer function

$$G_{12}(s) = \frac{Y_3 Y_4}{Y_5(Y_1 + Y_2 + Y_3 + Y_4) + Y_3 Y_4}$$

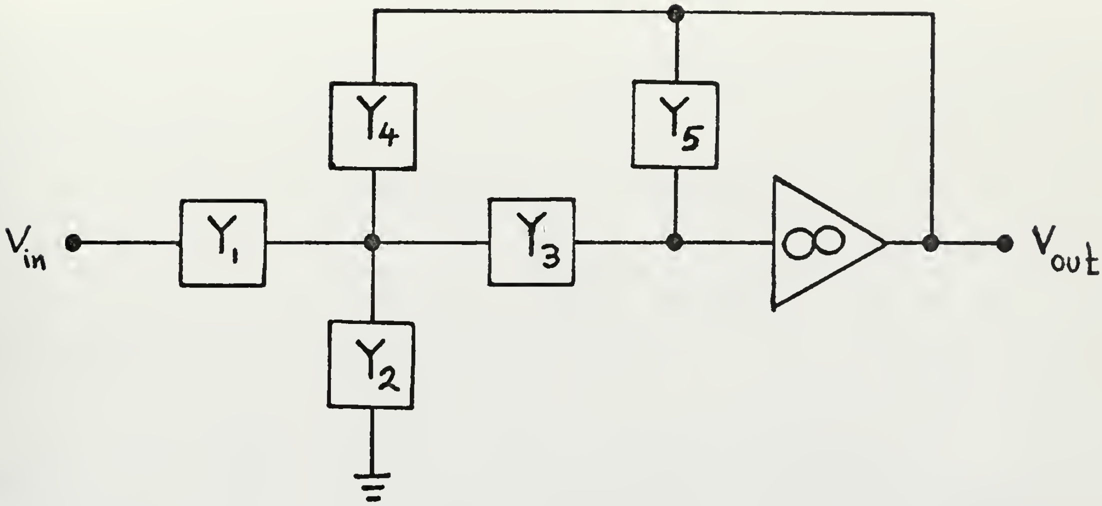


Figure 13. Brennan and Bridge's network.

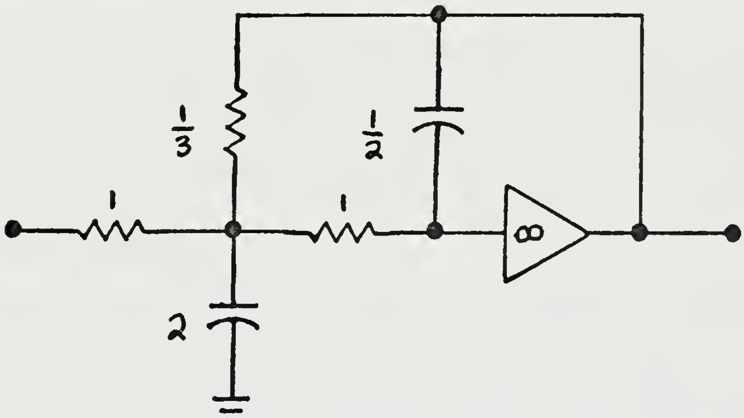


Figure 14. Realization of $G(s) = \dfrac{2}{s^2 + 3s + 3}$

By Brennan and Bridge's coefficient matching method [11] and by choosing 's suitably as

$$Y_1 = \frac{1}{R_1},\ Y_2 = C_2 s,\ Y_3 = \frac{1}{R_3},\ Y_4 = \frac{1}{R_4},\ Y_5 = C_5 s,$$

R's and C's can be related to the coefficients of the delay function as,

$$R_1 R_3 C_2 C_5 = 1 \qquad R_1/R_4 = 3$$

$$(R_1 + R_3 + \frac{R_1 R_3}{R_4}) C_5 = 3.$$

One solution to the equations gives

$$R_1 = 1,\ C_2 = 2,\ R_3 = .5,\ R_4 = 1/3,\ C_5 = 1.$$

Then, the circuit in Figure 13 changes to that of Figure 14. The transfer function is

$$G_{12}(s) = \frac{2}{s^2 + 3s + 2}$$

The constant "H" has been obtained as 2, which means the delay line has a pass band attenuation of 2/3. Therefore, this particular circuit is not a good choice.

1. <u>Chain network</u>

Flatter and wider frequency response and more linear phase characteristics are obtained when a higher order of truncation of the delay function is used. The "chain network" is a very convenient circuit for realizing higher order all-pole transfer functions [11]. The transfer function of an RC low pass filter in Figure 15.a is

$$G_{12}(s) = \frac{w_1}{s + w_1} \text{ where } w_1 = \frac{1}{R_1 C_1}.$$

Isolating time constants for two or more RC low pass filters

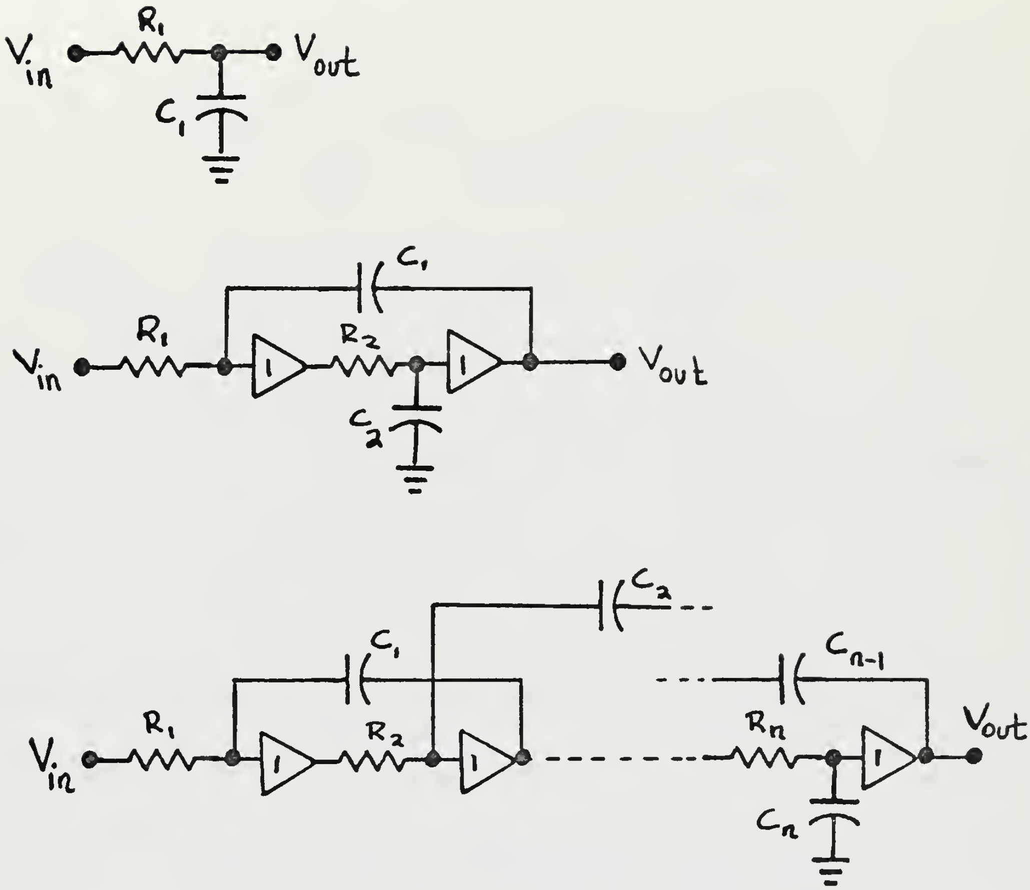


Figure 15. Development of RC chain network.

by using GUGA's,[1] the network in Figures 15.b and 15.c are obtained with transfer functions respectively.

$$G_{12}(s)_2 \frac{w_1 w_2}{s^2 + w_1 s + w_2} \quad \text{and}$$

$$G_{12}(s) = \frac{{}^{n}_{i=1} w_i}{s^n + w_1 s^{n-1} + w_1 w_e s^{n-2} + \cdots {}^{n}_{i=1} w_i}$$

where $w_i = \frac{1}{R_i C_i}$. An arbitrary voltage transfer ratio

$$G_{12}(s) = \frac{H}{s^n + a_1 s^{n-1} + a_2 s^{n-2} + \text{---} a_n}$$

is compared to the transfer function of the network in Figure 15.c, and by coefficient matching,

$$a_k = {}^{k}_{i=1} w_i \, (k = 1, 2, \ldots, n)$$

is obtained. The maximally flat delay function truncated at fifth order has a transfer function

$$G_{12}(s) = \frac{945}{s^5 + 15s^4 + 105s^3 + 420s^2 + 965s + 965}.$$

For this transfer function, w's of the chain network become,

$w_1 = 15$ ———————— $w_1 = 15$

$w_1 w_2 = 105$ ———————— $w_2 = 7$

$w_1 w_2 w_3 = 420$ ———————— $w_3 = 4$

$w_1 w_2 w_3 w_4 = 945$ ———————— $w_4 = 2.25$

$w_1 w_2 w_3 w_4 w_5 = 945$ ———————— $w_5 = 1.$

---

[1]GUGA stands for Grounded Unity Gain Amplifier as shown in Figure 16.

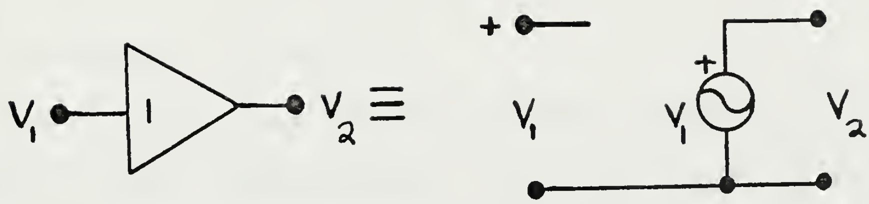


Figure 16. Equivalent circuit for GUGA

Assigning $C_i = 1.0$ $(i = 1,2,\ldots,5)$ yields,

$$R_1 = .066,\ R_2 = .142,\ R_3 = .25,\ R_4 = .66,\ R_5 = 1.$$

The realized network is shown in Figure 17. A delay line for five microseconds has been constructed by a chain network and the magnitude and phase characteristics of this line are given in Figures 18.a and 18.b. The response to a step input is given in Figure 19. There is a very close correlation between the step responses of the computer simulation and the actual model in Figures 9.e and 19.

One of the most severe problems encountered in active filters is their sensitivity. For this reason a sensitivity analysis was made for the fifth order chain network. The classical sensitivity function $S_x^N(p,x) = \frac{dN/N}{dx/x} = \frac{d(m/N)}{d(m,x)}$, where N is a network function and x the parameter varying, can be used to determine the normalized change in magnitude and phase of the network function N. The magnitude and phase sensitivities versus frequency, for the time constants $w_i = \frac{1}{R_i C_i}$ were calculated[1] and plots of those sensitivities are given in Figures 20.a through 20.e. (Horizontal axis frequency, vertical axis sensitivity).

2. <u>Hazony's network</u>

Another network which can be used as active delay line, as shown in Figure 21, is suggested by D. Hazony [12]. The unity gain amplifier in Figure 21 stands for a GUGA.

---

[1]Calculations are given in the Appendix.

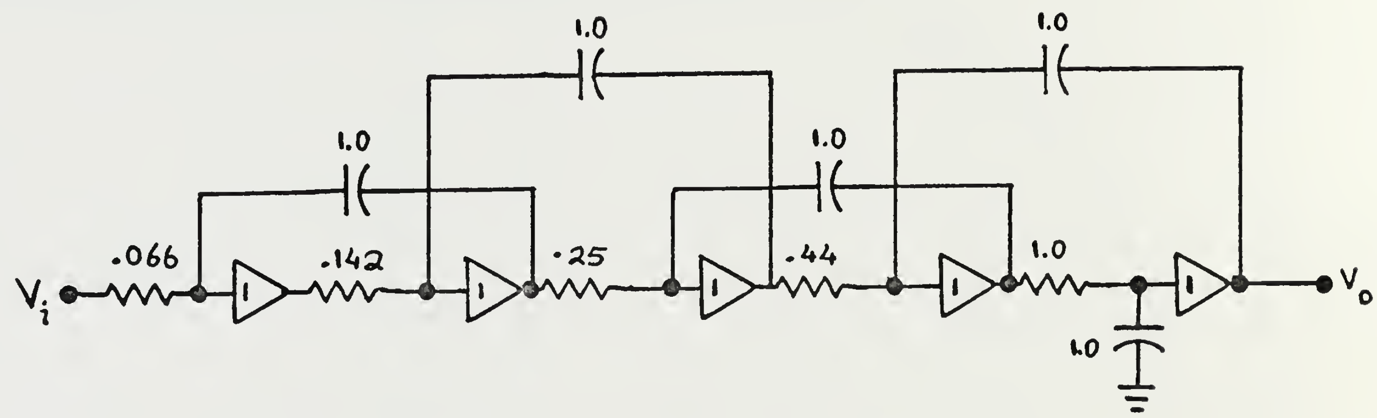


Figure 17. Fifth order chain network as delay line.

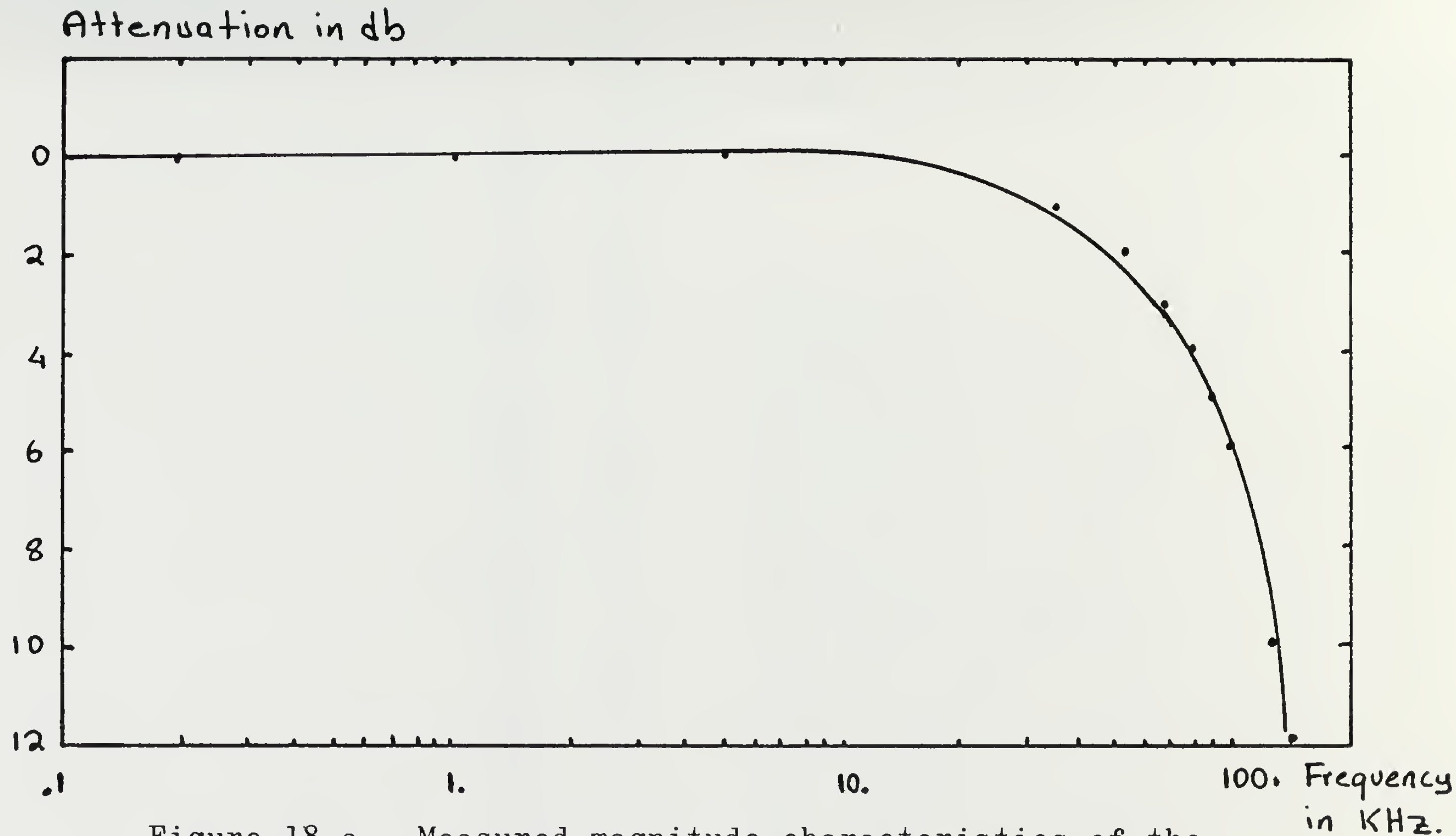


Figure 18.a. Measured magnitude characteristics of the network shown in Figure 17.

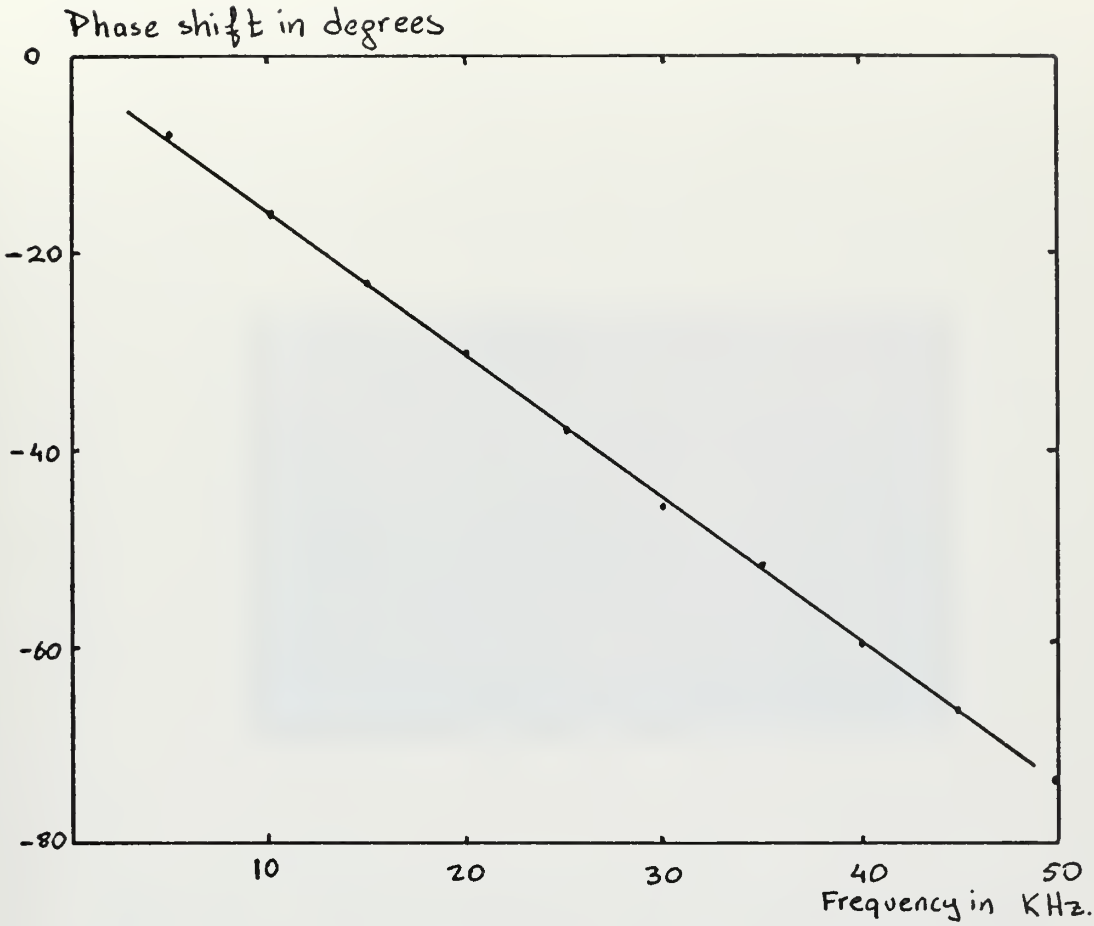


Figure 18.b. Measured phase characteristics of the network shown in Figure 17.

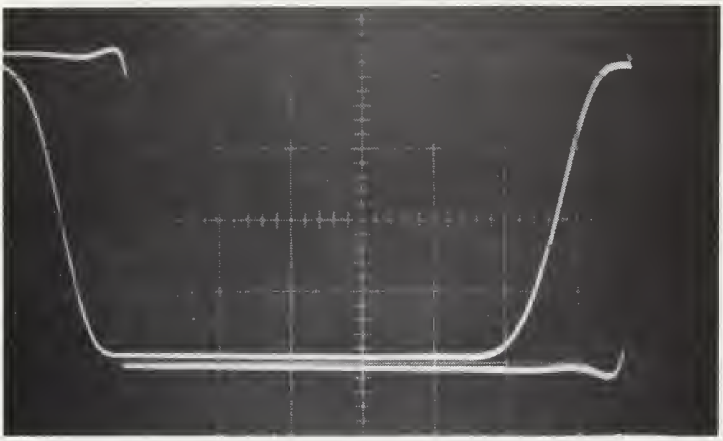
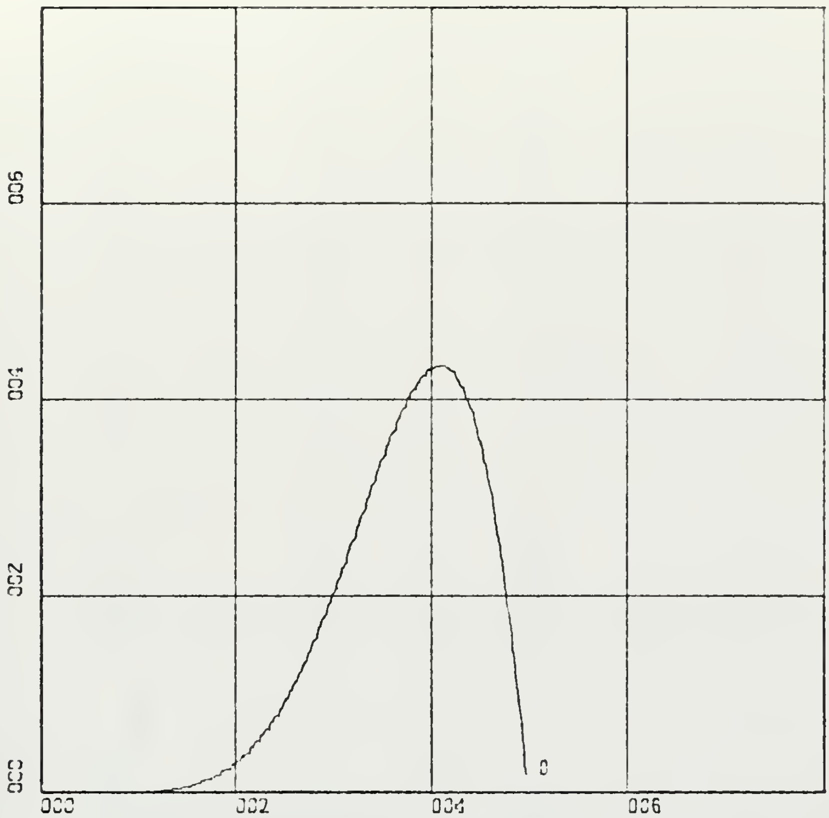
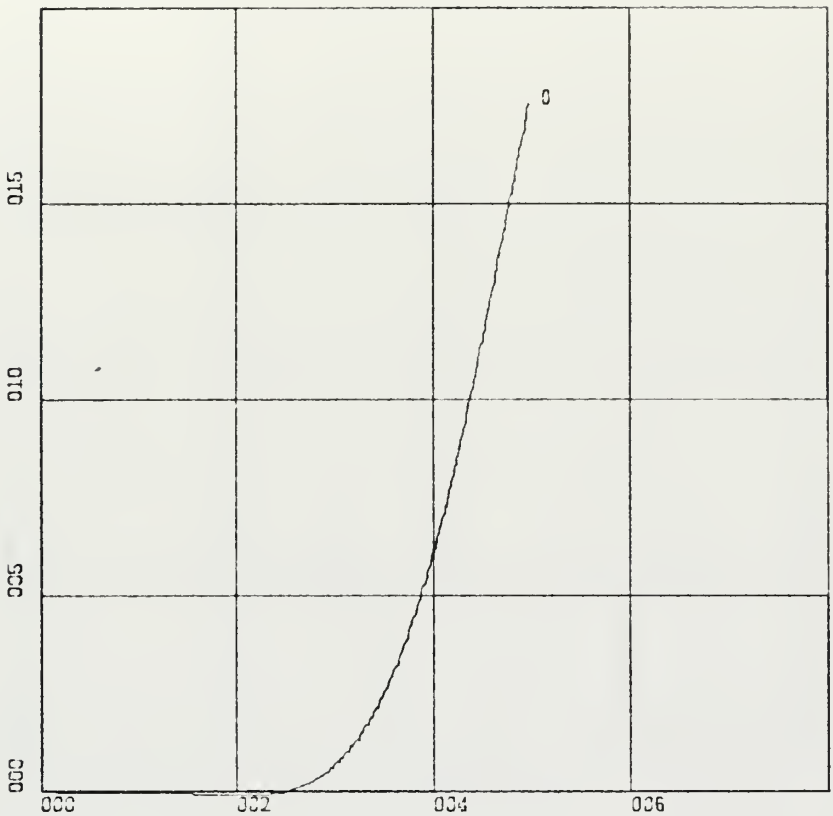


Figure 19. Step response of the constructed chain network delay line. (Horizontal scale 5 microsec./cm)



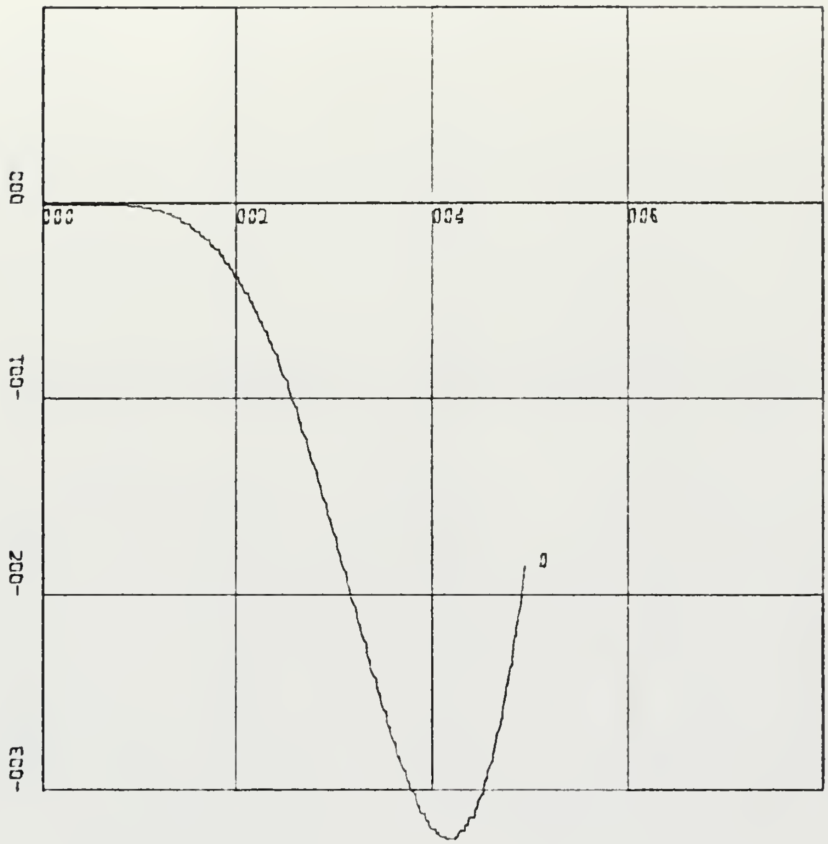
X-SCALE=2.00E-01 UNITS INCH.
Y-SCALE=2.00E-02 UNITS INCH.

Figure 20.a(1). Sensitivity of magnitude with respect to $w_1 = \frac{1}{R_1 C_1}$ for the transfer function given on page (39).



X-SCALE=2.00E-01 UNITS INCH.
Y-SCALE=5.00E-02 UNITS INCH.

Figure 20.a(2). Sensitivity of phase with respect to $w_1 = \frac{1}{R_1C_1}$ for the transfer function given on page (39).



X-SCALE=2.00E-01 UNITS INCH.
Y-SCALE=1.00E-01 UNITS INCH.

Figure 20.b(1). Sensitivity of magnitude with respect to $w_2 = \frac{1}{R_2 C_2}$ for the transfer function given on page (39).

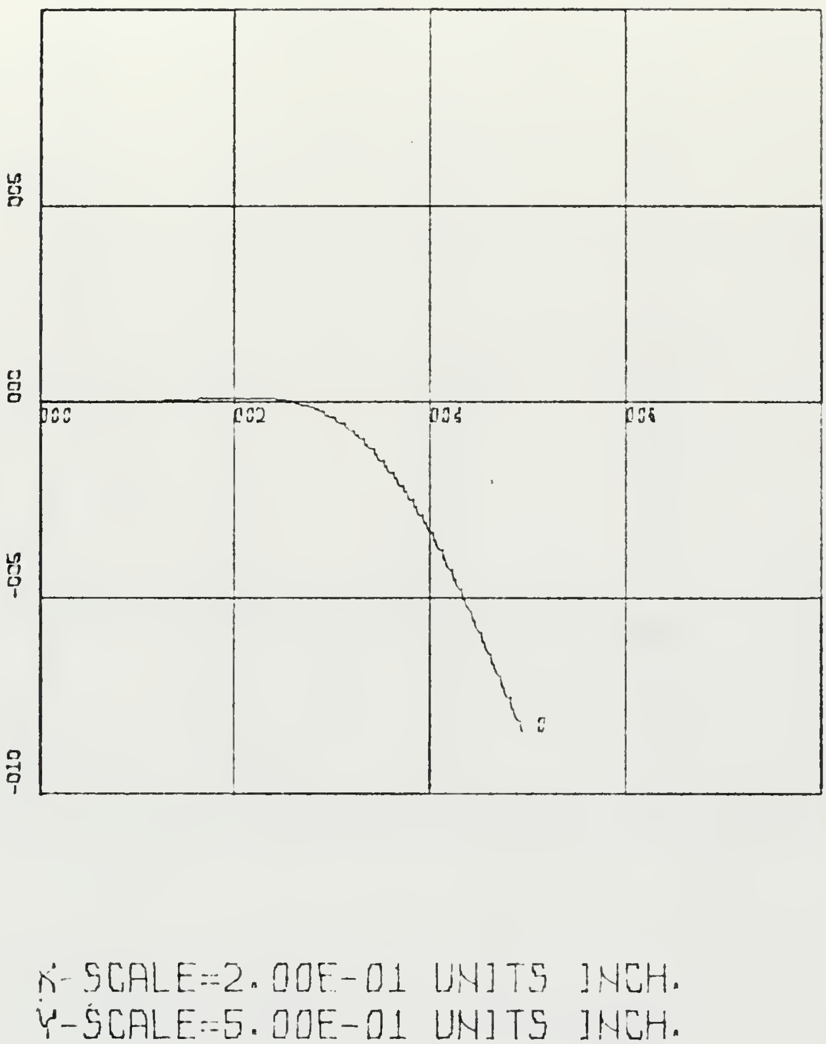
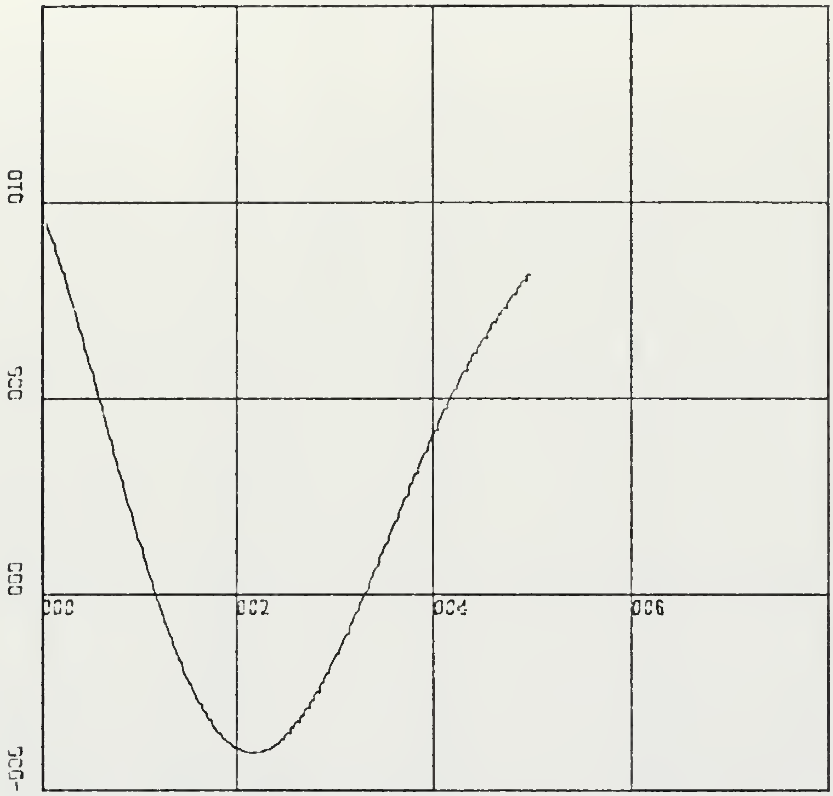
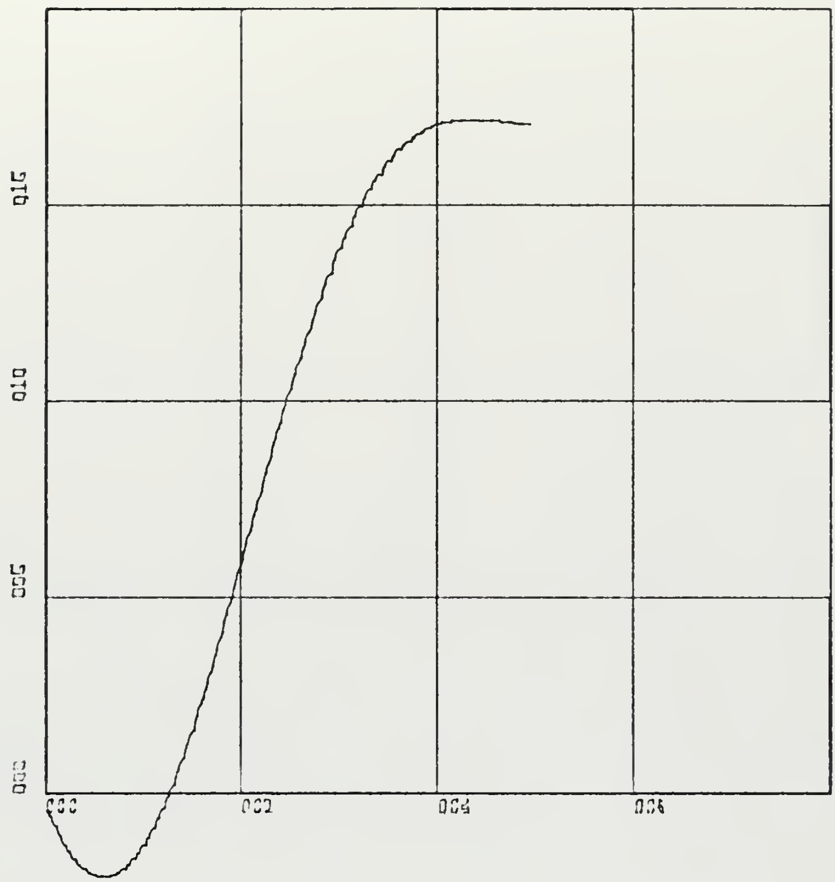


Figure 20.b(2). Sensitivity of phase with respect to $w_2 = \frac{1}{R_2C_2}$, for the transfer function given on page (39).



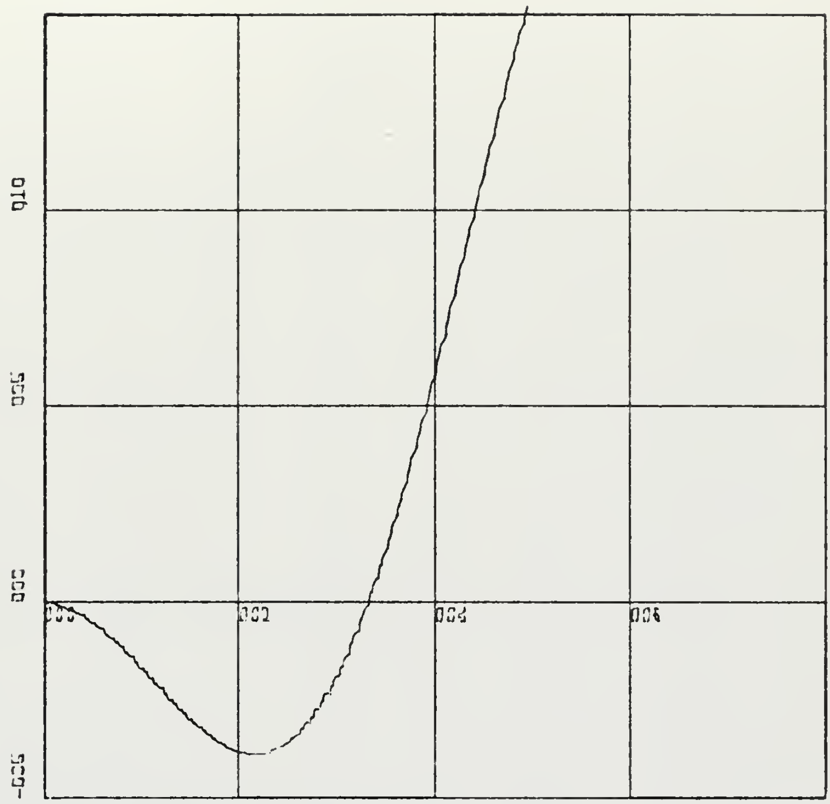
X-SCALE=2.00E-01 UNITS INCH.
Y-SCALE=5.00E-01 UNITS INCH.

Figure 20.c(1). Sensitivity of magnitude with respect to $w_3 = \frac{1}{R_3C_3}$ for the transfer function given on page (39).



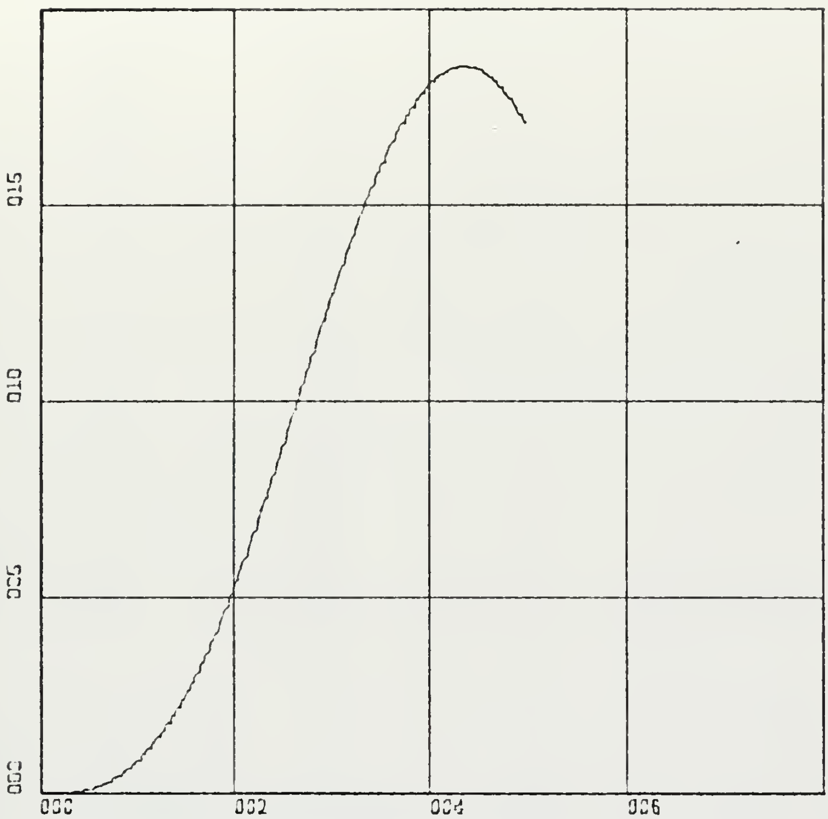
X-SCALE=2.00E-01 UNITS INCH.
Y-SCALE=5.00E-01 UNITS INCH.

Figure 20.c(2). Sensitivity of phase with respect to $w_3 = \frac{1}{R_3C_3}$ for the transfer function given on page (39).



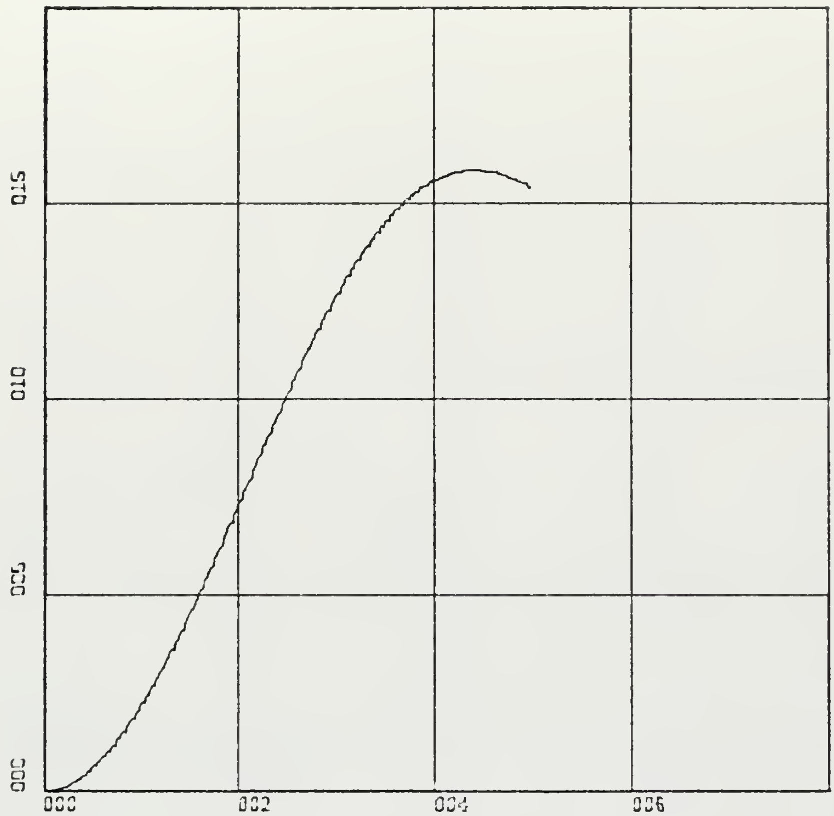
X-SCALE=2.00E-01 UNITS INCH.
Y-SCALE=5.00E-01 UNITS INCH.

Figure 20.d(1). Sensitivity of magnitude with respect to $w_4 = \frac{1}{R_4 C_4}$ for the transfer function given on page (39).



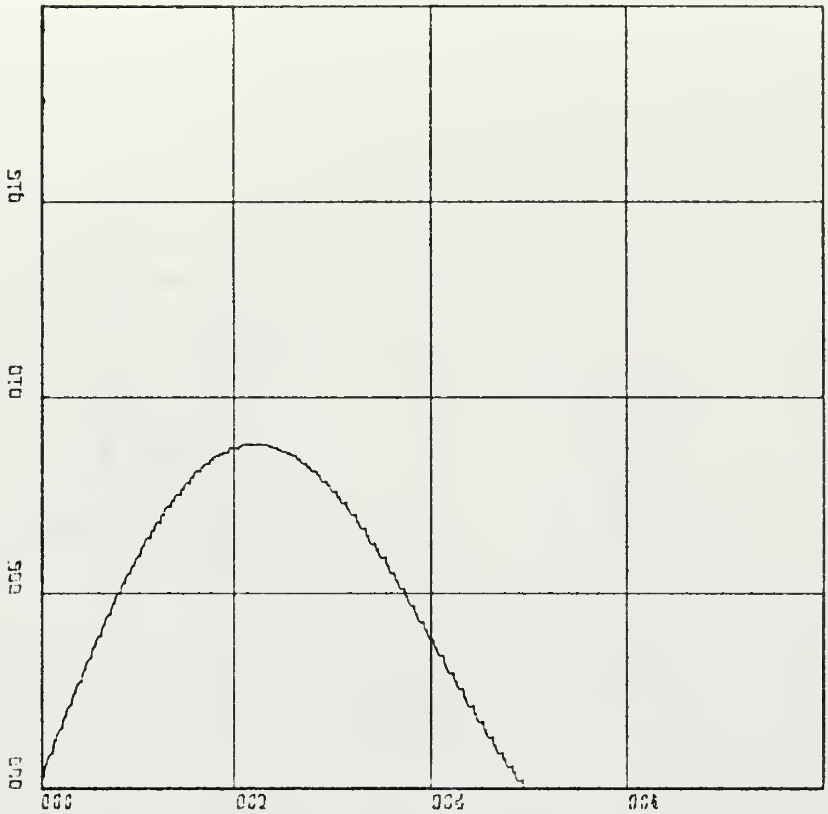
X-SCALE=2.00E-01 UNITS INCH.
Y-SCALE=5.00E-01 UNITS INCH.

Figure 20.d(2). Sensitivity of phase with respect to $w_4 = \frac{1}{R_4 C_4}$ for the transfer function, given on page (39).



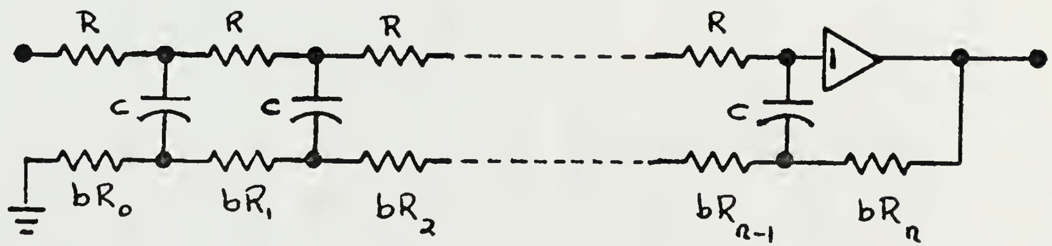
X-SCALE=2.00E-01 UNITS INCH.
Y-SCALE=5.00E-01 UNITS INCH.

Figure 20.e(1). Sensitivity of magnitude with respect to $w_5 = \frac{1}{R_5 C_5}$ for the transfer function given on page (39).



K-SCALE=2.00E-01 UNITS INCH.
Y-SCALE=5.00E-01 UNITS INCH.

Figure 20.e(2). Sensitivity of phase with respect to $w_5 = \frac{1}{R_5 C_5}$ for the transfer function given on page (39).



$$R = \frac{1}{\sqrt{n}} \qquad C = \frac{2}{\sqrt{n}}$$

Figure 21. Hazony's near canonic delay line.

The delay function e^s can be expanded by the binomial theorem as,

$$e^s \cong \left(1 + \frac{s}{n}\right)^n = 1 + nC_1\left(\frac{s}{n}\right) + nC_2\left(\frac{s}{n}\right)^2 + \cdots nC_n\left(\frac{s}{n}\right)^n$$

where

$$nC_i = \frac{n!}{(n-i)!\,i!} = \binom{n}{i}$$

The scaled resistors connected from the output terminal of GUGA are related to the binomial coefficients as,

$$R_i = \begin{cases} \dbinom{n}{\frac{n+i}{2}} & n+i \text{ is even} \\[2ex] \dbinom{n}{\frac{n+i+1}{2}} & n+i \text{ is odd.} \end{cases}$$

With the scaling constant "b," the scaled resistors are made considerably lower than those on the input side of the amplifier. These scaled resistors form a voltage divider and each resistor acts as an ideal voltage source. Choosing $n = 5$, a network as shown in Figure 22 is obtained. A delay line with delay time of 5 microseconds has been constructed by utilizing this network. The measured magnitude and phase versus frequency plots are given in Figures 23.a and 23.b and its response to a step input is given in Figure 24. For comparison purposes, the step responses of "chain network" and "Hazony's network" for fifth order approximation to a delay line with a delay time of 100 nanoseconds are shown in Figures 25.a and 25.b. The first pulses are inputs and delayed pulses are outputs.

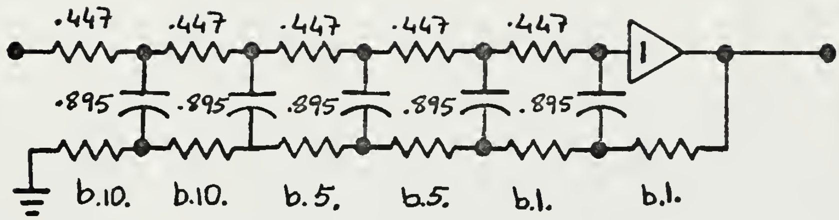


Figure 22. Hazony's fifth order delay line.

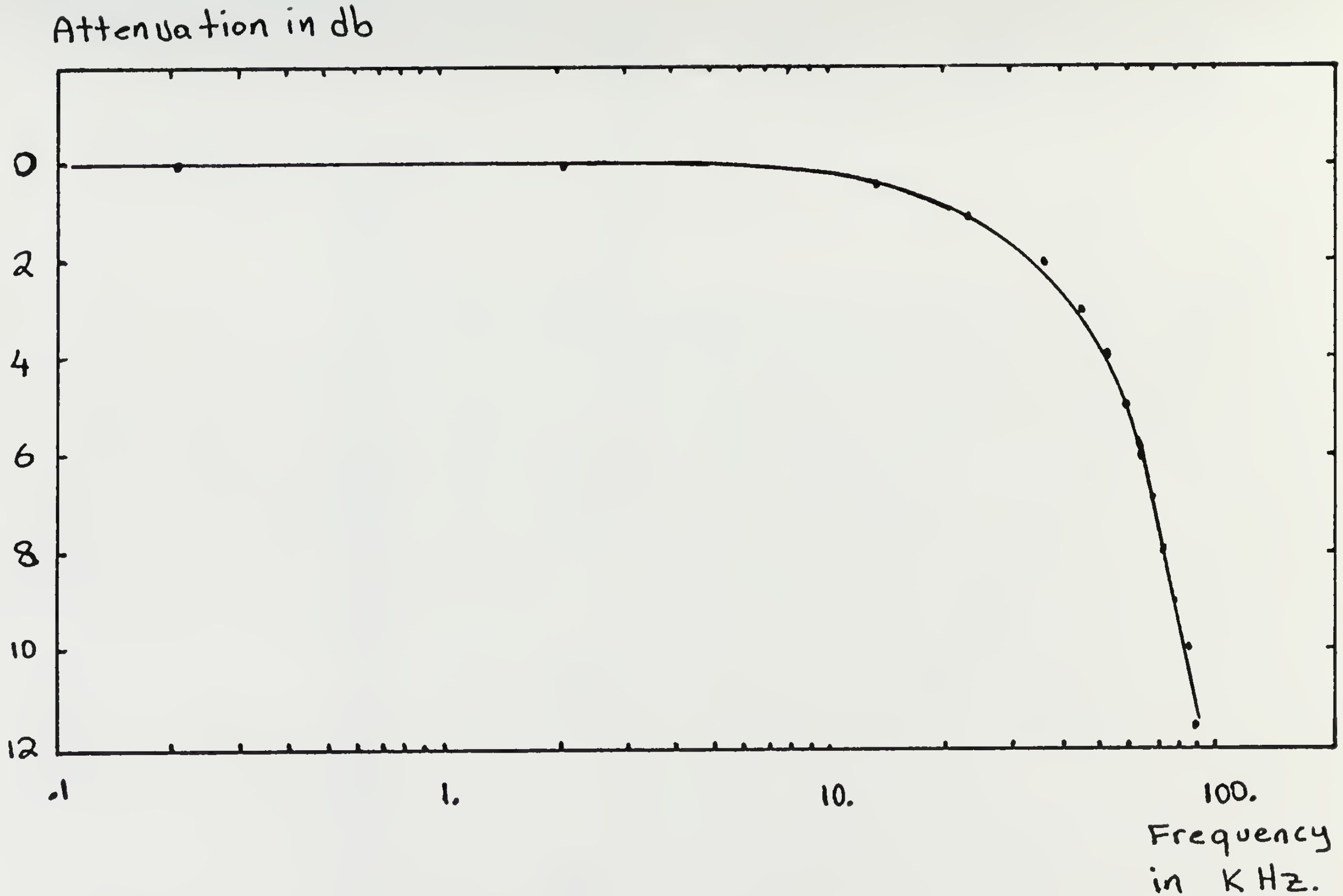


Figure 23.a. Measured magnitude characteristics of the network shown in Figure 22.

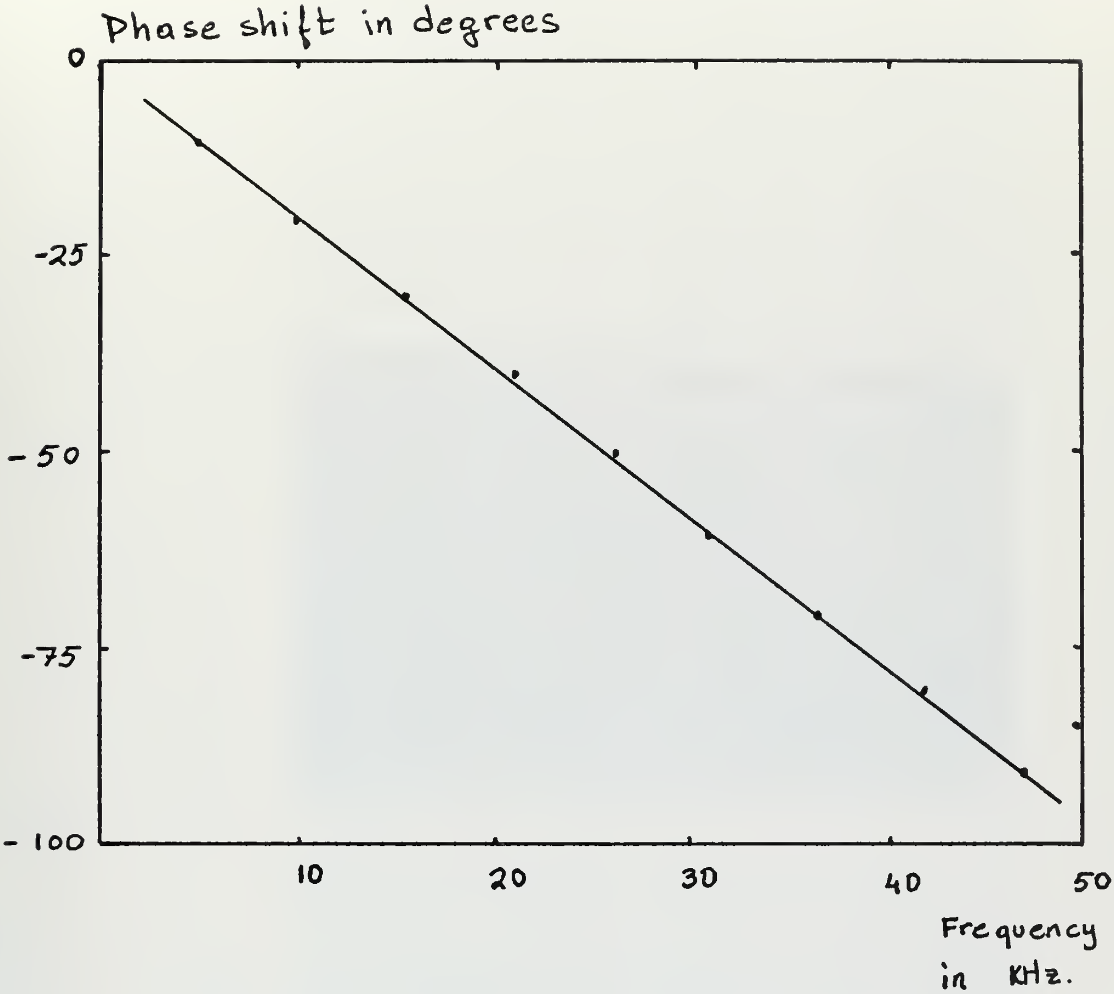


Figure 23.b. Measured phase characteristics of the network shown in Figure 22.

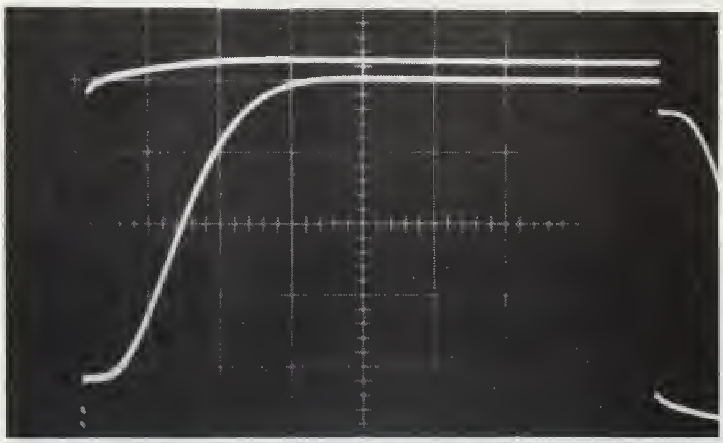


Figure 26. Response to a step input of Hazony's fifth order delay line.
(Horizontal scale 5 microseconds/cm)

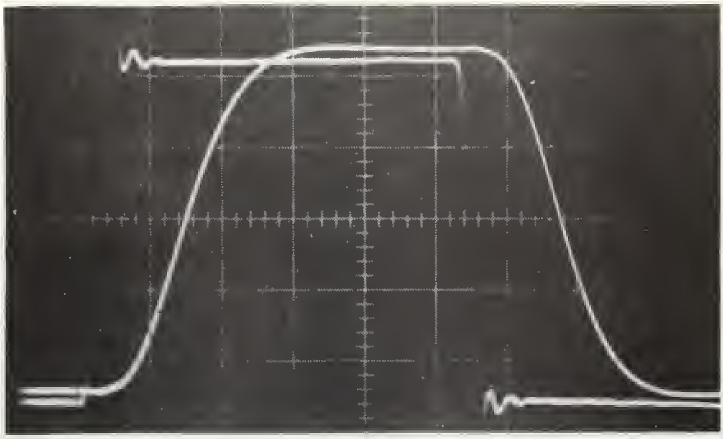


Figure 25.a. Response of chain network to step input.
(Horizontal scale .1 sec/cm)

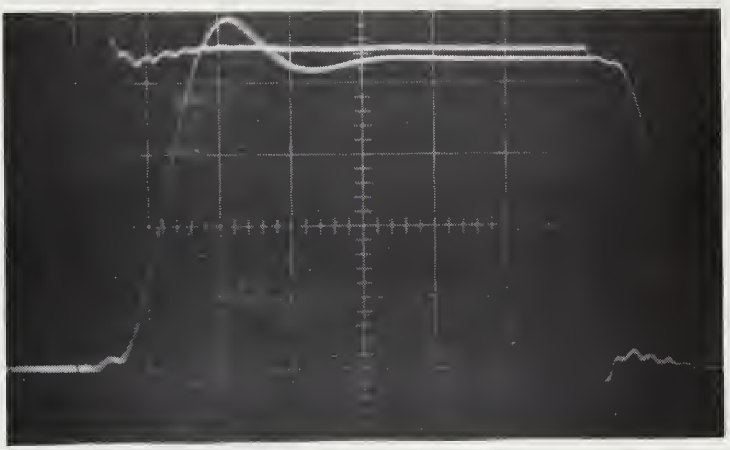


Figure 25.b. Response of Hazony's network to step input.
(Horizontal scale .1 sec/cm)
Both delay lines scaled to 100 nanoseconds.

IV. CONCLUSIONS

A simple add-on noise blanker to combat impulse type noise is feasible, provided that two problems can be solved effectively. First, detection of noise pulses in the presence of strong CW signals must be accomplished. This can be done by utilizing AND-gate type detectors. Second, obtaining a delay line with wide bandwidth and low loss. This delay line can be either a passive or an active one. Coaxial cables for delay lines take too much space and are lossy, although cheap. Lumped element lines with constant-k sections are easy to construct but yield too many sections, again too much volume and attenuation. The m-derived sections with toroidal cores have fewer sections with one core and one capacitor per section, but one of the inductors on the core becomes too small and hard to obtain precisely. This leads one to conclude that the only useful passive delay line for the blanker application is a distributed one.

The other alternative is an active delay line and this seems to be most promising because of its compactness and the possibility of having no loss. Because of required wide frequency response and large dynamic range, the active elements must be the lowest gain type possible, ideally, unity gain. The chain network and Hazony's network meet this requirement. The feasibility models of delay lines constructed are of fifth order approximation, giving quite low delay time over rise time ratios. This ratio can be improved (wider bandwidth

per unit delay time) by obtaining higher orders of approximation. With proper sampling of magnitude, tenth or higher order approximations can be realized. The chain network can also be used to realize a Butterworth filter which is maximally flat magnitude response function, therefore, has wider bandwidth than that of maximally flat delay function, although not very linear pulse characteristics. Since the main interest here is not in the shape of the pulse, this can be tolerated. Hazony's delay line has a slight advantage over the chain network in realizing higher orders of approximation, because of using only one unity gain amplifier, resulting in lower noise figure and less attenuation.

Wide use of integrated circuits can help obtain a high package density and ease of design. (A VHF blanker, operating on the same principles requires much greater complexity than an HF one, mainly due to limitations of gate and delay line circuits. A six diode gate using hot-carrier diodes [4] is promising for this application.)

V. RECOMMENDATIONS FOR FURTHER STUDY

The characteristics of active delay lines can be improved by realizing higher order transfer functions. The chain network and Hazony's network for higher order approximations should be constructed and tested. Also a comparison of the maximally-flat delay function and Butterworth function, phase and magnitude characteristics can be made by realizing both by the chain network. Pulse responses of these two can

be examined and a decision made as to which function is a better choice for the "noise blanker delay line."

The construction and evaluation of an entire noise blanker, using an active delay line would be the next step in continuing this study.

APPENDIX

The classical sensitivity is usually referred to as $S_x^N(p,x)$ and defined as [13],

$$S_x^N(p,x) = \frac{dN/N}{dx/x} = \frac{d(\ln N)}{d(\ln x)}$$

where N(p,x) is any network function and x any parameter. Let (p,x) be,

$$N(p,x) = \frac{Q(p,x)}{P(p,x)}.$$

Then it can be shown that,

$$S_x^N(p,x) = x\left(\frac{Q'}{Q} - \frac{P'}{P}\right),$$

where $Q' = \frac{\partial Q(p,x)}{\partial x}$ and $P' = \frac{\partial P(p,x)}{\partial x}$.

The classical sensitivity can be used to determine the variations in magnitude and phase of a network function caused by parameter variations. It can also be shown that the real part of $S_x^N(jw,x)$ specifies the normalized change in the magnitude of the network function and the imaginary part specifies the change in the phase function [14].

The network function for fifth order chain network is

$$G_{12}(s) = \frac{w_1w_2w_3w_4w_5}{s^5 + w_1s^4 + w_1w_2s^3 + w_1w_2w_3s^2 + w_1w_2w_3w_4s + w_1w_2w_3w_4w_5}.$$

The sensitivities for phase and magnitude of above function after substituting the coefficients of delay function[1] and $s = jw$ become,

---

[1] $G_{12}(s) = \dfrac{965}{s^5 + 15s^4 + 105s^3 + 420s^2 + 965s + 965}$

$$S^G_{w_1} = 1 - \frac{A^2 + BC}{A^2 + B^2} - j\,\frac{A(C - B)}{A^2 + B^2}\,,$$

$$S^G_{w_2} = 1 - \frac{AD + BC}{A^2 + B^2} - j\,\frac{AC - BD}{A^2 + B^2},$$

$$S^G_{w_3} = 1 - \frac{AC + BE}{A^2 + B^2} - j\,\frac{AE - BC}{A^2 + B^2},$$

$$S^G_{w_u} = 1 - 945\frac{A + Bw}{A^2 + B^2} - j945\frac{Aw - B}{A^2 + B^2} \quad \text{and}$$

$$S^G_{w_5} = 1 - 945\frac{A}{A^2 + B^2} + j945\frac{B}{A^2 + B^2},$$

where $A = 15w^4 - 420w^2 + 945$

$B = w^5 - 105w^3 + 945w$

$C = -105w^3 + 945w$

$D = -420w^2 + 945$ and

$E = 945w.$

The computed sensitivities for the above equations are plotted for $f = .01$ to $.5$ in Figures 20.a through 20.e.

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13. ABSTRACT

Atmospheric and man-made disturbances which result in noise pulses of random amplitude and randomly spaced in time have long been experienced in the communications field. Design features of an RF blanker for this type of noise are investigated. The specifications of functional blocks are established.

One of the functional blocks of an RF blanker is a linear delay line. The main features and design methods for passive delay lines are investigated. An active RC circuit to realize the delay function is proposed. This network and another are constructed and their characteristics obtained.

KEY WORDS	LINK A		LINK B		LINK C	
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